Electronic technology

principal contents

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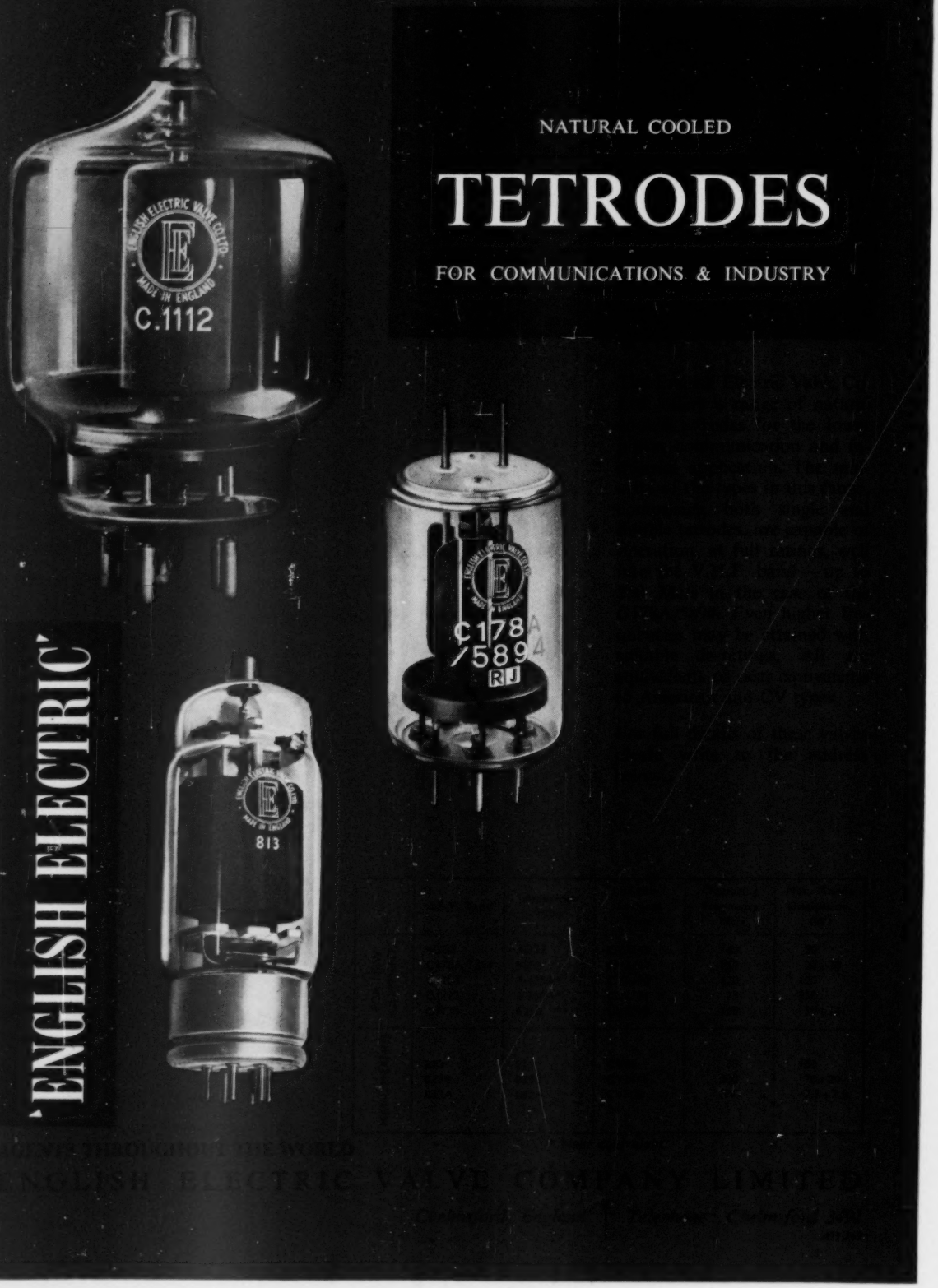
APRIL 1961 VOLUME 38 NUMBER 4



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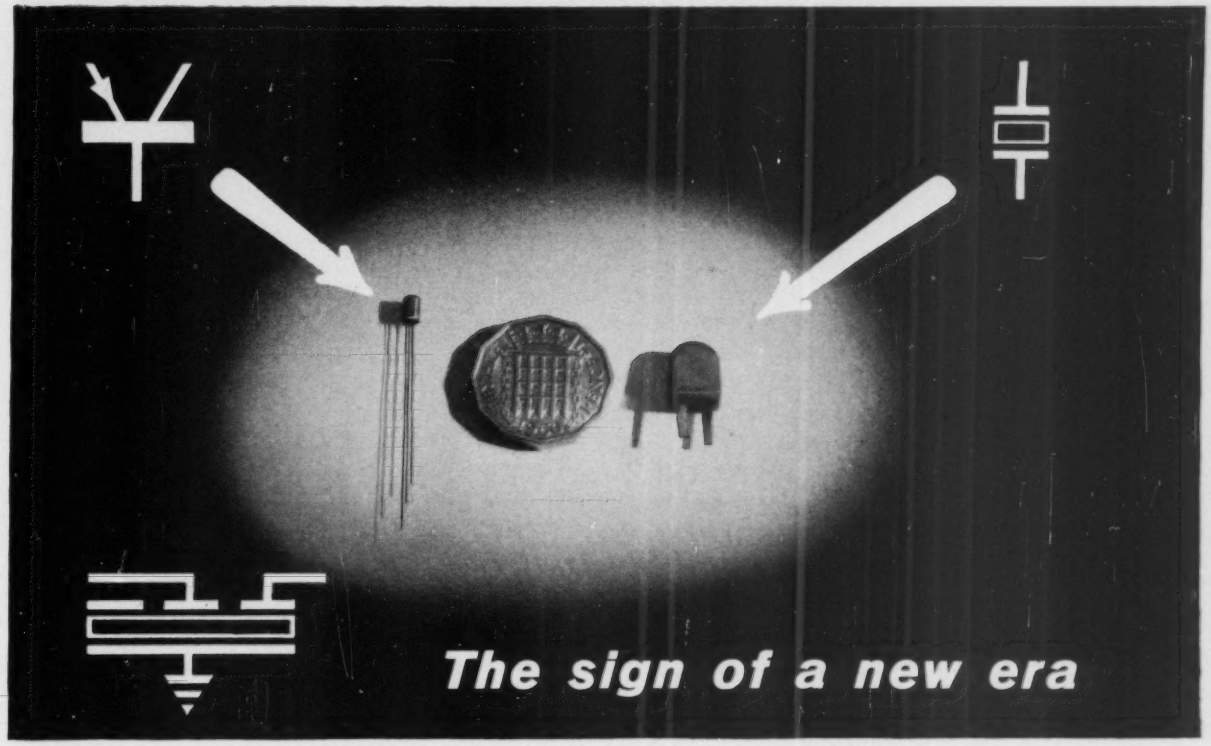

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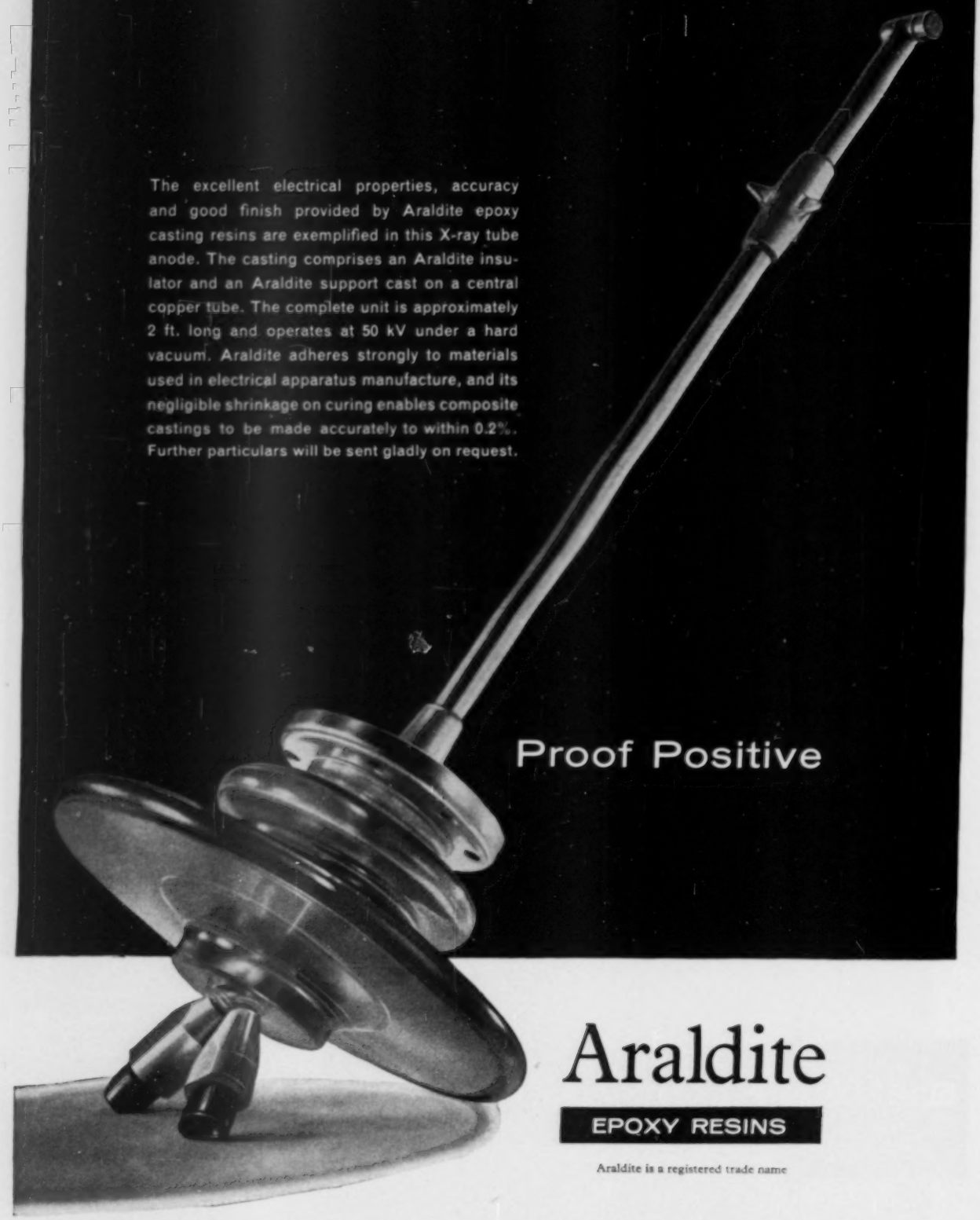

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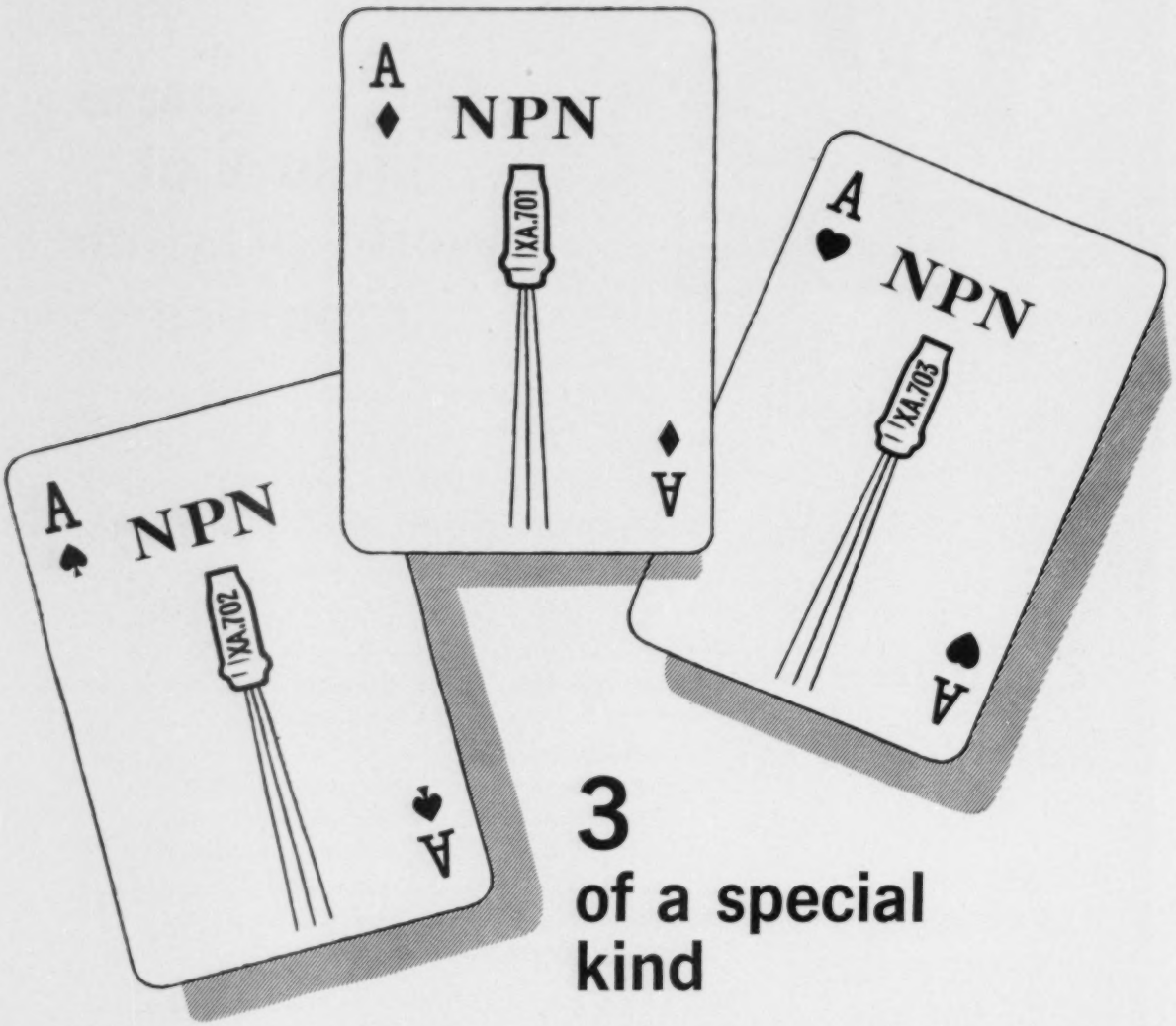

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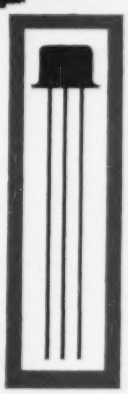

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CHARACTERISTICS.					
	ZT20	ZT21	ZT22	ZT23	Test Conditions
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Common Emitter A.C. Current gain h_{fe}	30*	60*	30*	45*	V_{CE}=6V I_C =1mA
Collector Saturation Voltage. Max. V_{CE} sat.	0·5V	0·5V	1·0V	1·0V	I_C =10mA I_B =2mA
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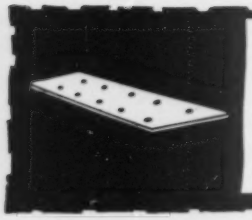

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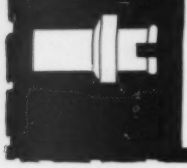

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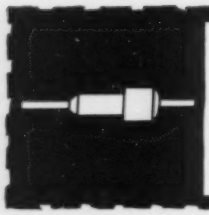

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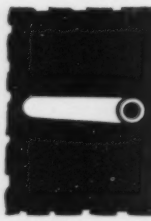

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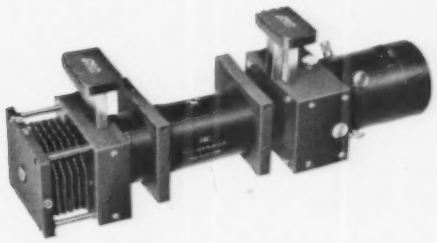

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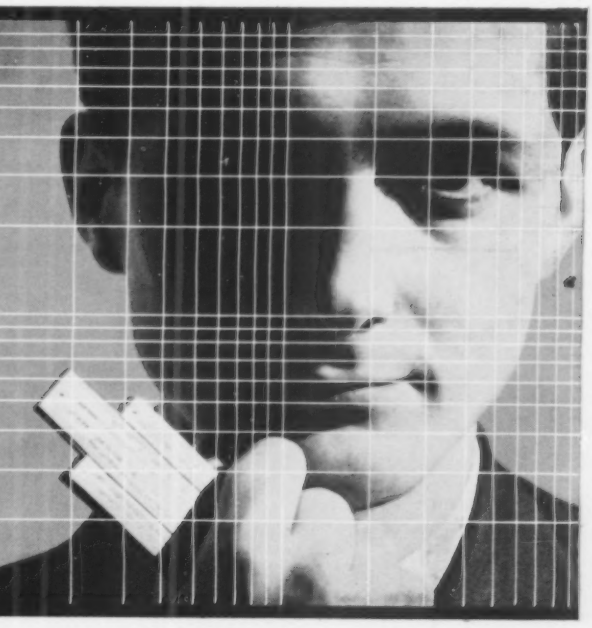
2·5 – 4·1	kMc/s	TWS6	1·0	20	30
2·7 – 3·5	kMc/s	TWS7	3·0	20	30
1·7 – 2·3	kMc/s	TWS10	18	27	
1·7 – 2·3	kMc/s	TWS1	30	32	
5·8 – 8·2	kMc/s	TWC5	11	39	30
7·0 – 11·5	kMc/s	TWX8	1·0	30	30



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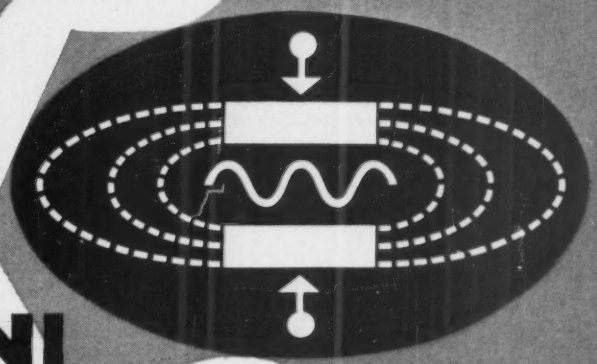

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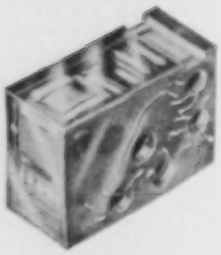

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An inductive probe is used in place of the capacitive probe for locating shorts between conductors. The oscillator will feed a signal to the conductor only up to the point where the short occurs and the probe will detect the electromagnetic field up to this point.

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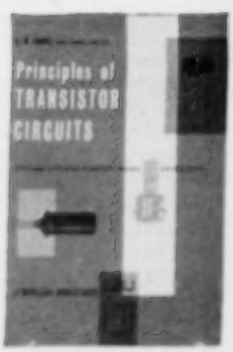

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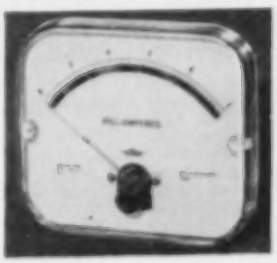

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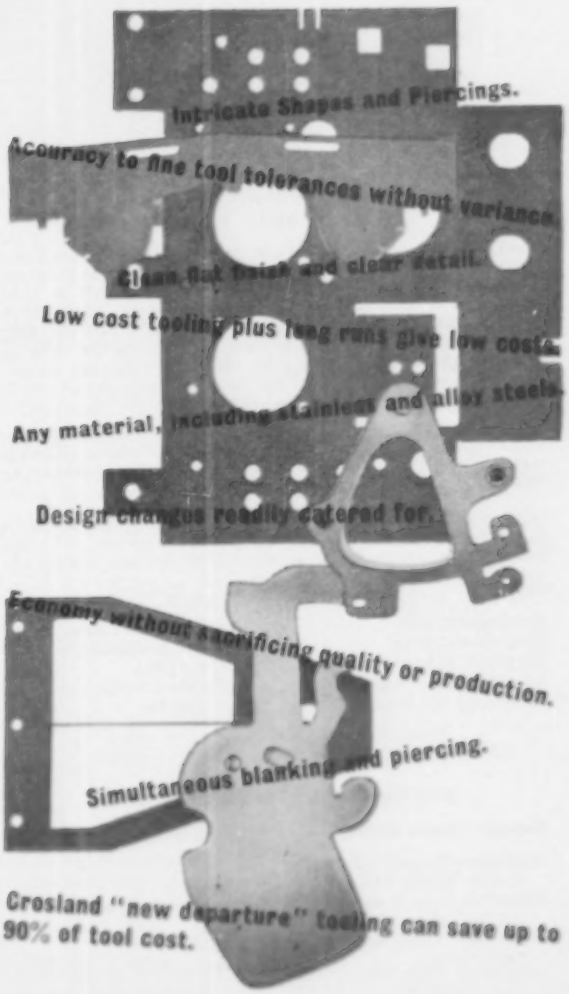
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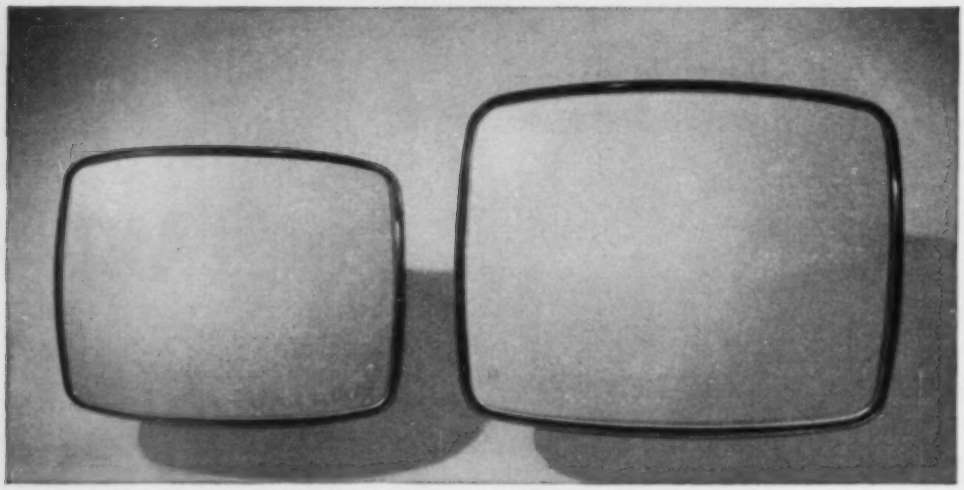

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The CME1901 and CME2301 are, respectively, 19in. and 23in. cathode ray tubes using magnetic deflection and electrostatic focus. The diagonal deflection angle of CME1901 is 114° and that of CME2301 is 110°. The shape of these tubes differs from the shape of conventional 110° tubes in that the face plates are more nearly rectangular. In addition the radii of curvature of the faces of the 19in. and 23in. tubes are greater than those of the 17in. and 21in. tubes. These changes result in a more pleasing presentation of the picture.

The external shape of the glass in the deflection region of these tubes is identical to that of conventional 110° tubes, enabling coils with conventional 110° internal mechanical contours to be used.

With equal values of final anode voltage, beam deflection in the CME1901 and CME2301 can be carried out with no more power consumption than in the CME1703 and CME2101.

GENERAL DETAILS

Rectangular face	Aluminised screen
Electrostatic focus	Silver activated phosphor
Magnetic deflection	Grey glass
Straight gun — non ion trap	External conductive coating
Heater for use in series chain	

Heater Current (amps)	I_h	0.3
Heater Voltage (volts)	V_h	12.6

TENTATIVE RATINGS AND DATA

Design Centre Ratings		**CME1901**	**CME2301**
Maximum Second and Fourth Anode Voltage (kV)	$V_{a2,a4\,(max)}$	17	17
Minimum Second and Fourth Anode Voltage (kV)	$V_{a2,a4\,(min)}$	14	15
Maximum Third Anode Voltage (volts)	$V_{a3\,(max)}$	±700	±700
Maximum First Anode Voltage (volts)	$V_{a1\,(max)}$	500	500
Maximum Heater to Cathode Voltage—Heater Negative d.c. (volts)	$V_{h\text{-}k\,(max)}$	180	180

Inter-electrode Capacitances (pF)		**CME1901**	**CME2301**
Cathode to All*	$C_{k\text{-}all}$	5	5
Grid to All*	$C_{g\text{-}all}$	8	8
Final Anode to External Conductive Coating (approx.)	$C_{a2,\,a4\text{-}M}$	1500	2000

*Inter-electrode capacitances including AEI "Clix" B8H holder VH68/81 (8 pin).

TYPICAL OPERATION

		CME1901	**CME2301**
Second and Fourth Anode Voltage (kV)	$V_{a2,\,a4}$	16	16-17
First Anode Voltage (volts)	V_{a1}	450	450
Third Anode Voltage for Focus-Mean (volts)	$V_{a3\,(av)}$	180	180
Grid Bias for cut-off of Raster (volts)		38 to 72	38 to 72
Average Peak to Peak Modulating Voltage for Modulation up to 350μA (volts)		34.5	34.5

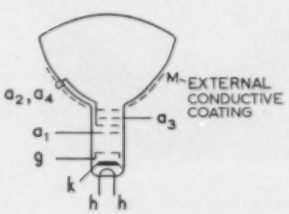
Note: All voltages given with respect to the cathode.

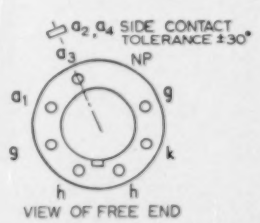
Maximum Dimensions (mm)	**CME1901**	**CME2301**
Overall Length	322	386
Face Diagonal	476†	598†
Face Width	420†	524†
Face Height	342†	422†
Neck Diameter	29.4	29.4

†The maximum dimension at the face seal may be 3.5 mm larger than this dimension but at any point around the seal the bulge will not protrude more than 2 mm.

Tube Weight (lb.)	**CME1901**	**CME2301**
Nett (approx.)	13.5	27

Side Contact: CT8 (Cavity) **Base:** B8H





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ELECTRONIC & RADIO ENGINEER

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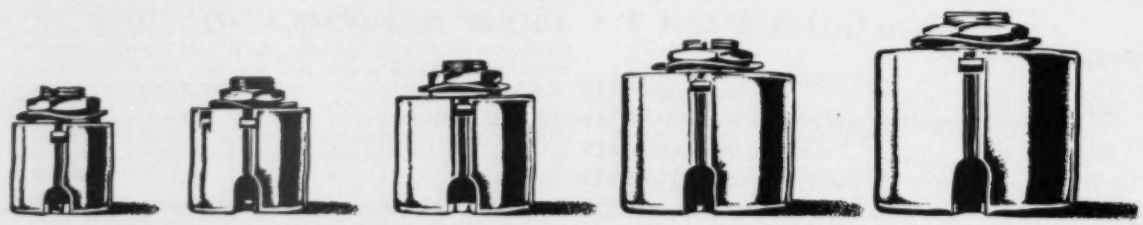

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APRIL 1961
VOLUME 38 NUMBER 4

Electronic technology

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"Wireless World"

THIS month our contemporary, *Wireless World*, celebrates its 50th year of publication with a special Jubilee issue in which is traced the history of wireless as recorded in its pages. This is nearly the history of wireless itself.

In April 1911, *The Marconigraph* was first published and was so successful that only two years later its title was changed to *Wireless World* in order that it might be in keeping with its greater circulation. Through the years *Wireless World* has always sought to record and explain technical developments in a field which has been continuously expanding. From the simple spark transmitter and magnetic detector of the early days, its coverage has expanded to all aspects of radio, radar, television and electronics, to say nothing of thermionics and semiconductors upon which they all depend.

Throughout its history *Wireless World* has been and still is outstanding for the attention paid to clarity of exposition in a subject which is not the easiest to write about. Believing, too, that true understanding comes as much from doing as from reading, *Wireless World* has never disdained to publish "how to make" articles. Thirty-five years ago improvements in valves, knowledge of how to construct low-loss r.f. inductances, and the development of the neutralized triode circuit coincided and for the first time made possible the attainment of useful and stable r.f. gain. Not content with describing and explaining all these things, *Wireless World* published constructional details of a receiver embodying them, the famous Everyman Four which has had an honoured place in the Science Museum for a good many years.

As time went on apparatus grew in complexity and, as it did, the practicability of publishing full constructional details gradually decreased. Nevertheless, high-quality audio amplifiers, superheterodynes with ganged tuning controls and even complete television receivers were described and, always, there were examples when these things were new developments.

Today *Wireless World* still maintains that practical outlook which is so helpful to the designer and still aims at clarity of exposition. We believe that this aim is being achieved to a remarkable degree.

Being born in 1923 we sometimes think of ourselves as an old journal, but *Wireless World* can claim 12 years' seniority and we are sure that our readers will all join us in wishing her in the future the same success that she has achieved in the last 50 years.

A Current Regulator and Sweep Mechanism

FOR THE ELECTROMAGNET IN PARAMAGNETIC RESONANCE EXPERIMENTS

By E. W. Collings *

SUMMARY. *Described here is a simple regulator and sweep mechanism for controlling the output of a d.c. generator. The sweep mechanism can vary the output current automatically over a wide range and the arrangement, which has a stability of* 1 *in* 4,000, *has been used as an electromagnet current supply in paramagnetic resonance experiments.*

The regulator is a modification of one previously described by Hedgcock and Hunt[1] which is suitable for regulating the current from either a battery bank or a d.c. generator. The present circuit, however, uses a push-pull d.c. amplifier in place of their single-ended one for amplifying the error signal and has the additional advantage of not requiring a reference battery. A current sweep mechanism has also been added.

Electronics

The electromagnet energizing current was obtained from a d.c. generator with separately-excited field coils. With the present equipment, for example, a field current of 100 mA gave an output of 4 A through a 16-Ω circuit which included a 4-in. Newport electromagnet. The field current was varied by altering the bias on a suitable number of power triodes (the 6AS7s in Fig. 1) and an amplified signal was introduced to the grids of these valves to stabilize the magnet-energizing current. A variable inductor (L, Fig. 1) adjusted the input voltage for optimum regulation, the correct setting of this component being readily found as the position at which it lost control of the magnet-energizing current.

A voltage proportional to the magnet-energizing current appears across a 2-Ω manganin resistor in series with the magnet (Fig. 2). This 'error signal' is amplified by the push-pull d.c. amplifier and applied to the grids of the series power triodes of the field-current power supply in such a sense as to achieve negative feedback. It can be seen that either P_1 or P_2 will vary the magnet-energizing current; when P_2 is an external Helipot for mechanical current sweep, P_1 is used to adjust the mid-sweep value of the current. The resistor bank has been designed so that mid-sweep current remains constant with change of the S_1-switch position.

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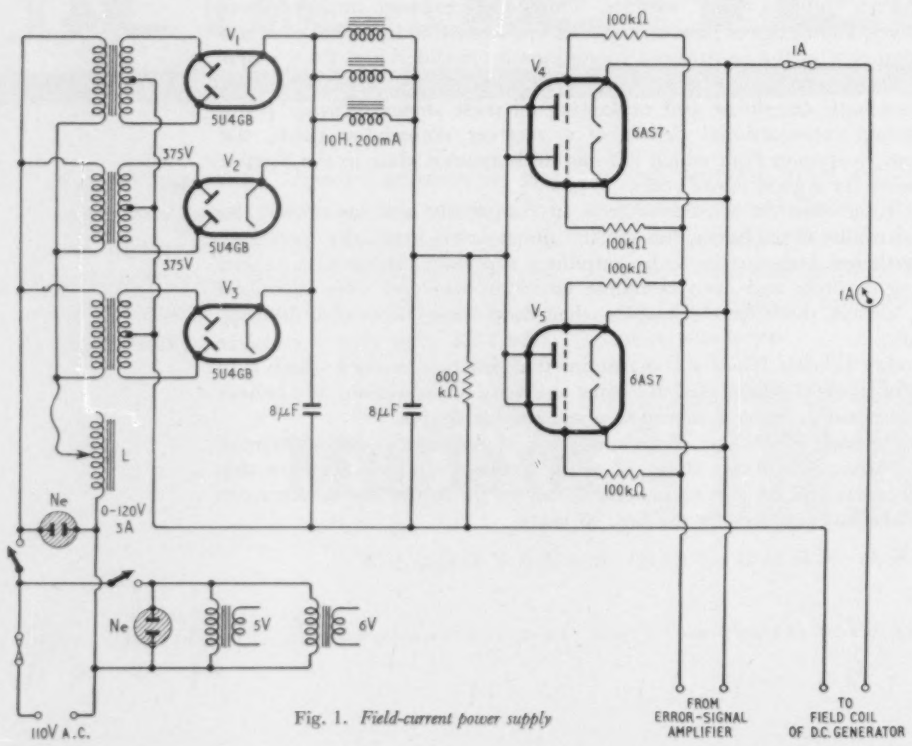


Fig. 1. *Field-current power supply*

Repetitive Sweep Mechanism

A 1-r.p.m. motor drives a ten-turn Helipot via a magnetic clutch. After ten revolutions of the drive shaft, a projecting screw head near the periphery of the larger gear of the 10 : 1 pair (Fig. 3) actuates a

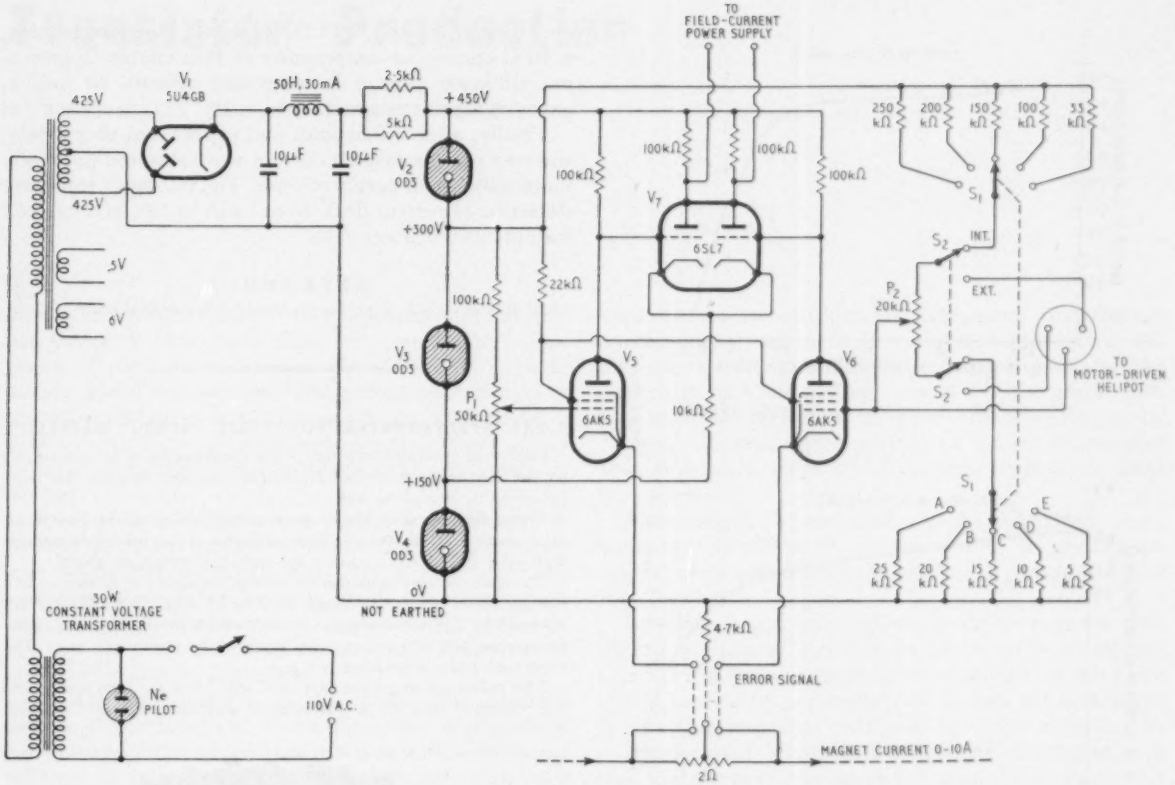


Fig. 2. *Error-signal amplifier*

microswitch and latching relay; the magnetic clutch then disengages and an auxiliary motor drives the Helipot back to its initial position. By varying the distance of separation of *two* such screws on the larger gear, the amplitude of the automatic repetitive sweep may be changed.

The sweep rate can be varied with the ganged switch S_1 (Fig. 2). From position A to position E the sweep rates were calculated to be in the ratio 1 : 2 : 3 : 4 : 5, the fastest sweep being 33 gauss/sec with a shaft speed of 1 r.p.m. For very slow sweeps of restricted amplitude, such as would be required when examining resonance lines of a few gauss in width, the 10-kΩ Helipot may be

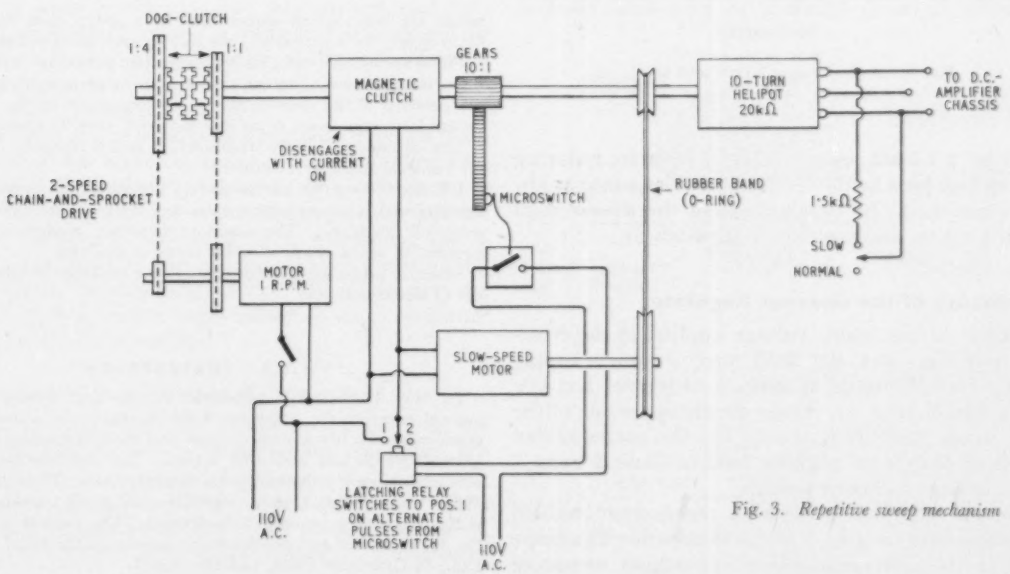


Fig. 3. *Repetitive sweep mechanism*

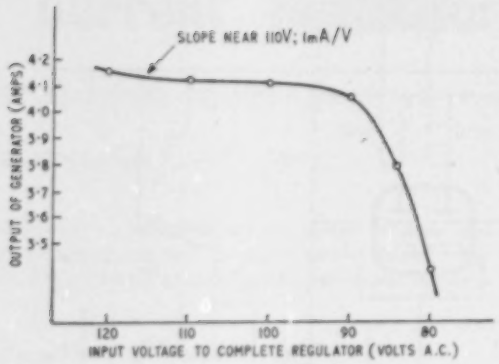


Fig. 4. *Stability against changes in mains voltage*

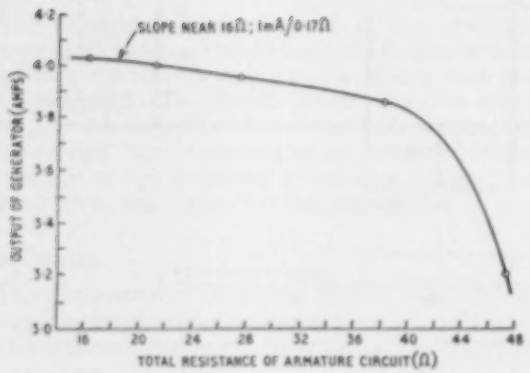


Fig. 5. *Stability against changes in circuit resistance*

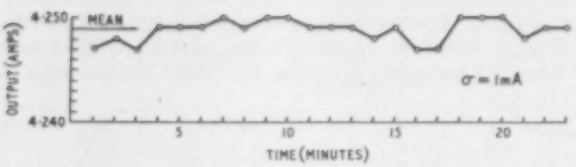


Fig. 6. *Drift of magnet current with time*

shunted by a 1·5-kΩ resistor (Fig. 3), thereby reducing the sweep rate by a factor 13. This enables a rate as low as 0·1 gauss/sec to be obtained using the slower shaft speed of $\frac{1}{4}$ r.p.m. and position A of switch S_1.

Performance of the Current Regulator

Variation of the mains voltage applied to the error-signal amplifier and the field-current power supply showed a rate of change of magnet current of 1 mA/V a.c. near 110 V (Fig. 4). From the characteristic of the magnet whose gap setting was $2\frac{1}{2}$ in., this corresponded to a rate of change of magnetic field of about $\frac{1}{2}$ gauss/1 per cent change in mains voltage.

A rheostat placed in series with the magnet enabled the characteristic of Fig. 5 to be obtained. The slope near 16 Ω (the total resistance of 5-Ω magnet, armature and wiring) is 1 mA/0·17 Ω. It would therefore require a 40°C change in temperature of that circuit to give a one-milliamp change in generator current, or half a gauss change in magnetic field.

Finally, after the magnet and supply had thoroughly warmed up, the magnet current was measured potentiometrically over a period of time. Fig. 6 shows a standard deviation of current drift (σ) of 1 mA in 4 A, or a current stability of 0·025 per cent.

REFERENCE

[1] F. T. Hedgcock and F. Hunt, *Rev. Sci. Instr.*, 1956, Vol. 27, p. 970.

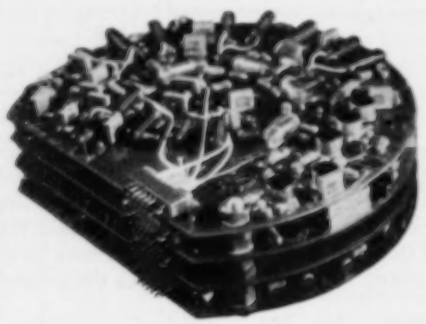
X-RAY SPECTROMETER FOR FIRST 'SCOUT' SATELLITE

Electronic equipment for one of the experiments to be carried out by the first Anglo-American Scout space-research satellite has been produced by Bristol Aircraft.

The equipment is an X-ray spectrometer which will be used in an experiment to obtain data for the correlation of ionospheric behaviour with solar X-radiation outside the earth's atmospheric shield.

The spectrometer measures the rate of incidence of X-rays in five energy bands within the range of 2 to 13 angstroms. X-rays are detected by a proportional gas counter which produces pulses whose heights are proportional to the energies of the incident X-rays. The minimum pulse separation is 1 μsec.

The pulses are amplified (about 72 dB), they are then sorted into five energy bands by a pulse-height discriminator, and the five



bands are interrogated sequentially by a gate; each band being interrogated for a period of 1 sec during each cycle of operation.

Pulses occurring during each interrogation period are accumulated in a 15-stage binary counter and, at the completion of each count, the contents of the counter are transferred direct to the telemetry system for transmission or on to a magnetic tape for storage.

The equipment is fully transistorized and is designed to operate on less than 100 mW of power.

UK Scout 1 consists basically of a 23-in. diam. 10-in. deep central cylinder with a hemisphere at each end, the lower one carrying four solar-cell 'paddles'. The interior electronic equipment is constructed in $5\frac{1}{2}$-in. diam. circular cards or modules.

When in orbit, the UK Scout satellite is expected to have a useful life of about one year.

"A.E.I. ENGINEERING"

In May 1960, A.E.I. published for the first time a technical journal covering the activities of the Company as a whole. The event marked a break with the past and the disappearance of house journals which had been well known. The first four issues of the new journal were published as an interim edition. With the January 1961, Vol. 1, No. 1 issue, monthly publication commences and *A.E.I. Engineering* becomes fully fledged. The journal is edited by Dr. D. C. Miller and published by Associated Electrical Industries Ltd., 33 Grosvenor Place, London, S.W.1.

Transistor Production

A NEW FACTORY AT SOUTHAMPTON

Progress of development of semiconductor devices has probably been more rapid than any other single group of components used in the electronics industry today. From the discovery and partial appreciation of transistor properties in 1948, a multitude of devices have evolved which are not only replacing thermionic valves in large quantities but also many other conventional devices.

Production techniques have played no small part in making transistors realizable devices. This is particularly true when one considers that only a few years ago transistors were being made in small quantities under laboratory-type conditions by highly-skilled staff and yet now they are produced in very large quantities in factories by normal operators.

An excellent example of this is the new Mullard factory at Southampton, where the initial impression is one of surprise due to the apparent ease with which some millions of semiconductors are being produced each month. On line after line, assemblies are flowing without any obvious difficulty. In particular, one notices that the operating-theatre environment, which was once so obvious a part of transistor production, has disappeared. The reason for this is very simple; transistor technology has advanced to the stage where it is known which processes require a controlled atmosphere and which processes can be carried out in a normal atmosphere. Therefore, instead of very large 'clean-rooms', in which operators wear protective clothing similar to that of the surgeon, and extremely complicated atmospheric filtering is needed, small bench-mounted cabinets are used only where it is necessary to have a controlled atmosphere for assembly. Access to the inside of the glass-fronted cabinets is gained through two apertures in which arm-length rubber gloves are sealed.

The success of any mass-production unit depends upon the ability of that unit to quantity-produce economically. In the case of transistor production, this is made more difficult by the small size of the basic elements and the resulting wastage in producing these elements. Fig. 1 indicates the sizes of the parts of various transistors. Other factors contributing to the difficulties are the required purities of the basic materials and cleanliness demanded during assembly.

Germanium-Alloy Junction Transistors

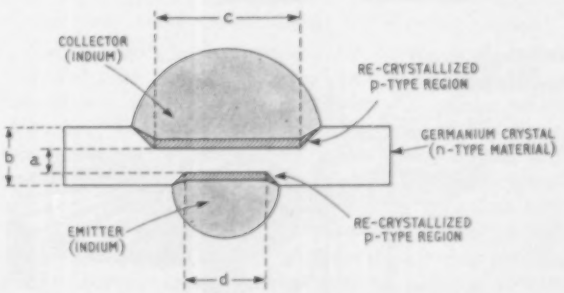
The basic materials used in the construction of the active element of a germanium-alloy junction are germanium, indium and antimony.

The germanium is initially melted down into crude ingots and zone-refined until the impurity level is reduced to 1 part in 10^8. Zone-refining, a universally-used method for purifying semiconductor materials, is a technique whereby a molten zone is progressively swept through a rod of the crude material carrying the impurities with the molten zone to the end of the rod. One standard method to achieve this in practice is by carrying the crude germanium in a quartz or graphite boat through a series of r.f. heating coils in an inert atmosphere.

After refining, a controlled amount of impurity (e.g., antimony) is added to the germanium to provide an ultimate controlled impurity level of 1 part in 10^6 and it is 'levelled'.

Levelling is one name given to the process for producing a uniform distribution of impurity along the crystal. It is somewhat similar to zone-refining but more closely controlled, particularly, as is often the case, when it combines levelling with making single crystals. In straightforward levelling, a single r.f. heating coil is used and the required impurity is loaded into the heated liquid zone. The concentration of the added impurity in the molten zone is such that, as the zone is swept through the ingot, small amounts of the impurity are delivered into the germanium in a uniform manner.

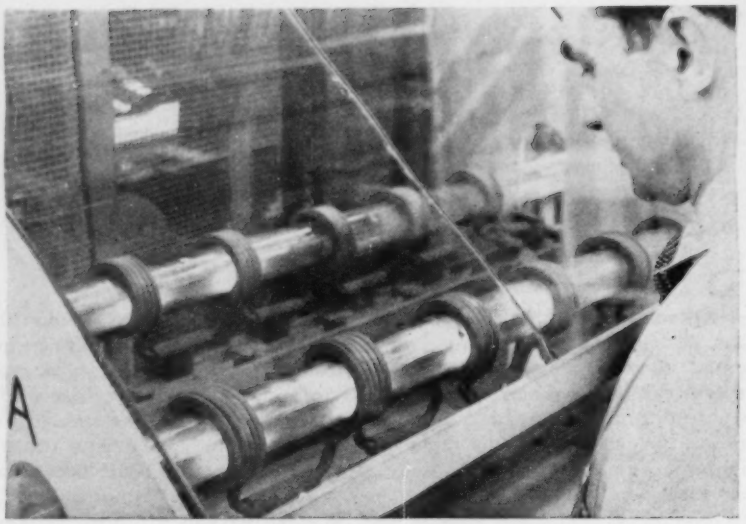
The structurally crude (polycrystalline) ingot of germanium is made into a single crystal by sowing a perfect single crystal seed with the required lattice orientation into one end of the germanium carrying boat and passing the boat very slowly, and at a very constant rate, through a single r.f. heating coil in an inert atmosphere. At the start, the single crystal seed melts a little and, as the boat proceeds through the



TRANSISTOR TYPE	TYPICAL DIMENSIONS (MICRONS)			
	a	b	c	d
R.F.	10	50	400	500
L.F.	25	150	1,100	700
HIGH-POWER	40	170	2,000	1,200

Fig. 1. *Shows a cross-section of a transistor and typical dimensions for various types*

heating coil, the germanium is melted and 'grows' on to the seed, taking up the same lattice orientation as the seed. During this process, almost any disturbance such as change of temperature, change of feed-rate, or mechanical vibration, will result in irregularities occurring in the final crystal. Major irregularities are almost immediately noticed, as they are visible on the surface of the crystal. When these occur, either the whole rod of germanium is re-processed or the acceptable part is cut off and used, while the remaining part is re-processed. Minor irregularities or faults occur in all crystals but these are checked on a sample basis by etching individual slices and microscopically or photographically inspecting the samples.



Close-up of the 6-coil germanium zero-refining unit

From the bars of single-crystal germanium (approximately 14 in. long), slices are cut with diamond-impregnated wheels to a prescribed thickness, a typical figure being 300 μ. The wastage of material is approximately 50%, since the width of the cut is comparable with the thickness of the slice of crystal. Following cutting, the slices are lapped, polished and cut into rectangular pieces (dice) from 1 mm square up to 5 mm square and, typically, 250 μ thick. Dicing is achieved by sticking the slice of crystal to a small piece of tape and scribing the surface in an automatic machine with 13 staggered diamond cutters and passing the crystal under a small steel roller to crack it into the scribed dice. The dice or crystals are then sorted by hand, chemically etched to a particular thickness and, to provide the required surface finish, cleaned and gauged to a predetermined thickness.

For the gauging of individual crystals, Mullard have developed and built a completely automatic machine which operates at a rate of 2,500 crystals/hr, gauges to an accuracy of $\pm$ 2 μ and segregates crystals into twenty thickness groups about a pre-set mean. This comprises a vibrator-driven container with a raised helical ridge around the inside perimeter, feeding the crystals one at a time between feeler rods, a detector circuit, and a servo system which controls the rotation of a segmented tray, according to the measured thickness of each crystal (Fig. 2). In operation, the crystals are loaded into the container and the vibration drives them in an

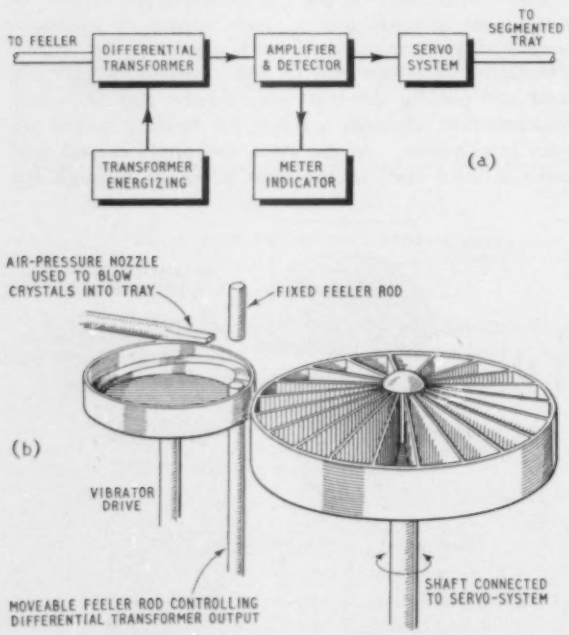


Fig. 2. *The block diagram (a) indicates the arrangement of the complete system while the sketch (b) shows the arrangement of the basic mechanical parts of the crystal gauging machine*

Fig. 3. *The two basic shapes of base tabs used in transistor production*

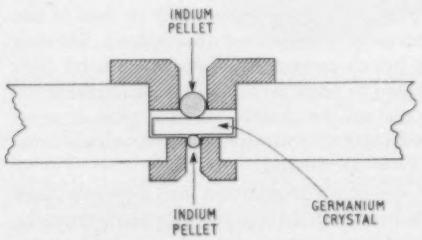
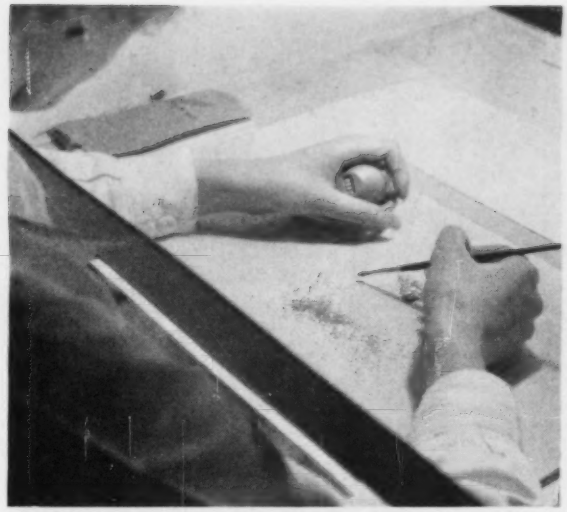
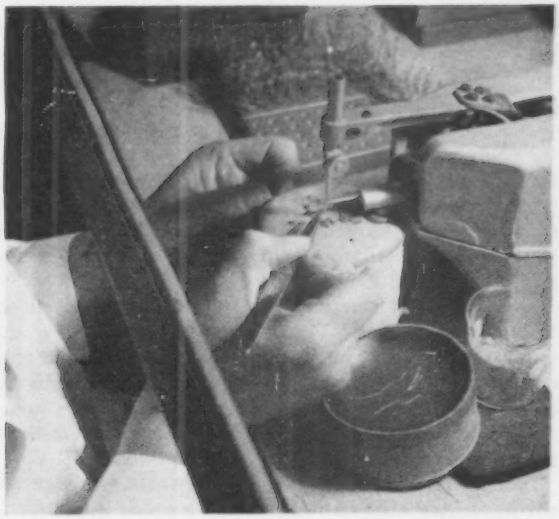


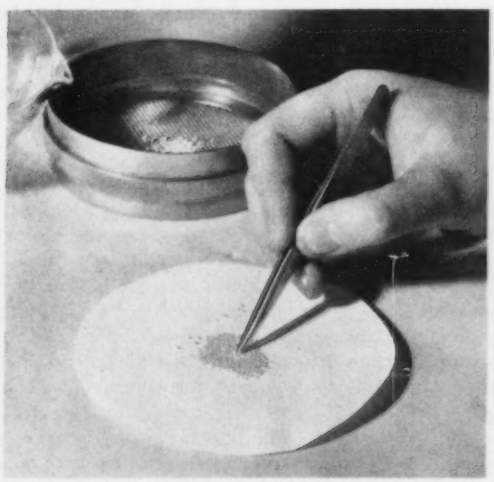
Fig. 4. *Shows one section of a ten-section jig which is used to hold the crystals and locate the indium pellets during the alloying operation*





(Left) Counting and sorting the germanium dice; although automatic methods have been tried these have so far failed, primarily because present automatic methods cannot efficiently detect square dice with corners missing. (Right) Under a glass cover an alloyed transistor assembly is being welded to glass beads

orderly convoy up the helical ridge on to a moveable feeler rod which lifts the crystal and presses it against a fixed feeler rod. The moveable feeler disturbs the magnetic circuit of the energized differential transformer to which it is connected and provides a change of output from the transformer which is proportional to the gap between the feeler rods and, therefore, proportional to the thickness of the crystal being measured. This output is amplified, detected and fed simultaneously to an indicating meter and a servo system, which is arranged to operate the drive to the segmented tray until the appropriate section is opposite the feeler rods. The feelers are released and a puff of air from a nearby nozzle blows the crystal into the tray. Crystals which are thicker than required are re-etched to reduce the thickness and re-measured.



Sorting and inspecting the indium pellets for the emitters and collectors of transistors

Other materials and component parts are prepared and made in parallel with the production of germanium crystals. One of these is the indium pellets for the transistor emitter and collector. Indium, which is 99·99% pure, is rolled and punched into discs, the dimensions depending upon the actual device to be made but, typical for an l.f. device, the diameter of the discs would be 0·6 mm for the emitter and 1 mm for the collector. Later, the discs of indium are formed into spheres and chemically or ultrasonically cleaned. During the process of making the indium pellets, the material and tools are kept scrupulously clean and the dimension tolerances are tightly controlled so that the ultimate weight of each pellet is within $\pm$ 3 or 4% of a pre-determined value.

Before the alloying operation, the base contact tab, which is a punched nickel tag with a hole in the centre (Fig. 3), is cleaned and tinned so that it can be soldered to the germanium crystal during the alloying process.

For the alloying operation, the germanium crystal with the base contact is loaded into a jig, designed to take 10 sets of component parts, with the indium pellets (Fig. 4). The loading of these jigs is semi-automatic, inasmuch as the component parts (i.e., crystal and pellets) are dispensed into the jigs automatically, ten at a time, while the jigs are on a moving conveyor belt. The jigs are then loaded on to a second conveyor belt and passed through two furnaces for the alloying process, the first furnace being used to 'wet' one pellet of indium and the second to 'wet' the other pellet and alloy both with the germanium. Alloying is carried out in a reducing atmosphere containing hydrogen at a temperature of about 500 °C which is controlled $\pm$ 2 °C for a period, depending upon the device, from 10 to 30 min. The overall time cycle for alloying (cold to cold, through a controlled rising-and-falling temperature gradient) is of the order of 30 min. to 3 hr. The higher the temperature in the furnace, the greater the depth of penetration

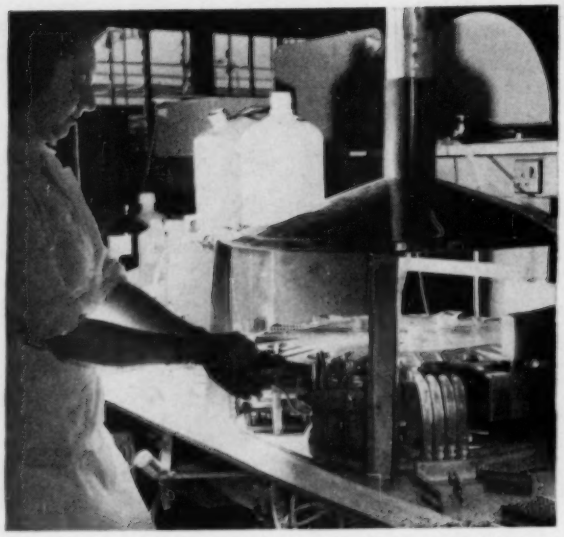
of the indium and, therefore, the smaller the distance between the inside faces of the alloy regions (base width), which effects the highest operating frequency of the end product.

It is interesting to note that up to the alloying stage the component parts and assembly are exactly the same for all transistors in one major group such as the OC44/45 series. It is only during the alloying process that the final characteristics of a particular batch of transistors are determined by accurate control of the process time and temperature cycle.

Following the alloying, the junction is mounted on to a header or foot. This is done by spot-welding the base contact tab and jumper wires to the header contacts (Fig. 5); the jumper wires are connected to the collector and emitter by 'soldering' with heated reducing gas or, in the case of high-power devices, with a small soldering iron.

The junction is then electrolytically etched to remove the unwanted alloy around the collector and emitter to clean the surface of the germanium. This is carried out by mounting the header assemblies into 10-unit jigs and loading these jigs into revolving gantries in a rotary etching bath, the contacts of the headers being connected to one side of a d.c. supply, while the other side is connected to an electrode in the electrolyte.

After the final etching, the units are washed in de-ionized water and baked at a slightly elevated temperature in an oven with a controlled oxygen atmosphere to oxydize the surfaces of the transistor element. At this stage the transistors are working devices and are pre-tested before encapsulation.



Jigs containing ten transistor assemblies being put into a rotary etching and spray washing unit

The type of encapsulation used for most low-power devices is a glass bulb partially filled with an inert filler, such as silicone grease with heat-conducting additives. Here the header assemblies are loaded into a machine which automatically loads and flame-seals the glass bulbs to the header beads.

For the final electrical test of the transistors, semi-automatic pre-programmed machines are used. The operator connects the leads into a quick-release socket and uni-selectors connect the appropriate potentials, signals and measuring circuits and, in a few seconds, panel lights clearly indicate to the operator that the unit under test is to be dropped into a particular shoot. The same machine is also designed for selecting matched sets of different types of transistor. From the final test the transistors are taken for type numbering, re-test and packing.

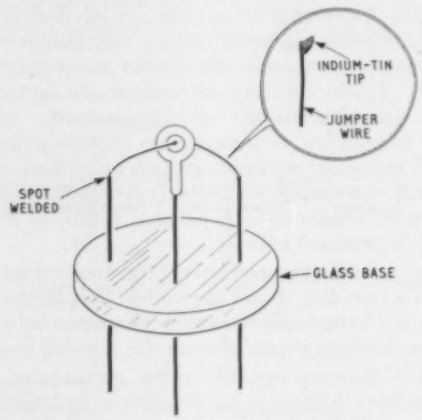


Fig. 5. *The glass base or bead of a transistor with the tab assembly mounted on the contact wires*

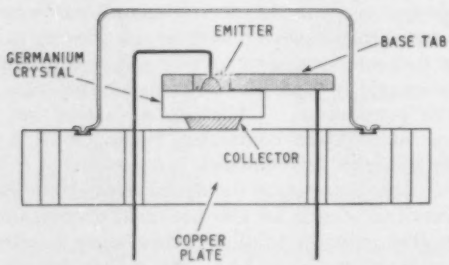


Fig. 6. *A cross-sectional sketch of a high-power transistor*

Germanium Power Transistors

The production techniques for the basic elements for high- and low-power transistors are the same. However, the elements for high-power devices are usually larger and the assembly of the elements into the mounting case is modified to improve heat conduction. First, the base contact tab is larger; secondly, most of the hemisphere of indium forming the collector is sheared off to provide a larger contact area and it is soft-soldered to a large copper plate which forms the mounting base for the complete transistor. Finally, large-diameter copper connecting wire is used throughout. Fig. 6 shows a cross-section of a typical high-power transistor and illustrates the assembly technique employed to improve heat conduction.

ROOM AT THE TOP

You can, of course, see what it is all going to be about this month—the conduction of electricity in metals. Everybody knows why metals conduct electricity; it is because the Fermi level happens to come within the conduction band, and because this is itself unfilled, so that there is room for more energetic electrons in the band above the Fermi level. I hope you are satisfied with this. You ought to be, because there is a mathematical justification of it all complete, as far as it goes, with the usual idealizations and necessary simplifying assumptions, which you will find in any of the recent advanced textbooks. But I also hope that some of you are dissatisfied with the way it is explained in the more elementary (and even popular) books into which the inescapable ideas of modern physics seem to filter.

These perpetual schematic diagrams of bands—cross-hatched and speckled, ruled and scrawled, and largely vacant space anyhow—do they really mean anything to the reader? And the potential diagram that goes hoppity-hoppity-hop, which must have been copied regularly from an original source of twenty years ago that I have been unable to trace, where does that come in? At the risk of being tedious, and laying myself open to obvious criticism, I should like to examine these two pictures rather closely.

Every picture, as one of the old masters used to say, tells a story. Like General Stanley did in *The Pirates of Penzance*, very often. But it is usually an emotional rather than a reasoned document. Cold reason at an early age prevented me from following smooth injunctions to use Anti-antimacassar Oil (though I fell for the little girl in the advertisement, nits and all); and from putting on my best velveteens to blow bubbles. Nowadays the emotional appeal is stronger and deeper, and our new masters have us drifting anxiously through the prestige-ridden day into the starvation-haunted night; while, sad to say, our nominal and elected guardians cannot even exploit this simple technique convincingly. I am thinking bitterly of the advertisement "Come Back to Teaching", with its dreary little pencil-sucking puppet, which suggests only that what the country really needs is a determined Education of Ministry; and contrasting it with the effective sales-talk about organization-man careers for the "right sort of chap", by which the best young graduates are ruthlessly, selfishly, and unnecessarily deflected away from the urgent calls of the teaching profession. But I must get back to the story behind the two pictures I started on. And not loiter on the job. For the horrid thought has struck me that the time is not far distant when, like all my fellows, I shall have to give up writing; for the simple reason that the new generation will not be able to read.

Energy Levels in the Free Atom

We must first see how it is that in a real solid the Fermi level, which was calculated for a gas of free electrons at absolute zero, can play an important part. In order to discuss this, you remember, we had to remove the whole of every atom in the solid except for the one electron it contributed, and it won't be long before we have to put them back. We can forget about the absolute-zero part, and also the effect of ordinary temperatures on the motions of the replaced atoms, at least for the moment; for we agreed that, as compared with the Fermi-level energies of the order of several electron volts, thermal-agitation energies at ordinary temperatures must be trivial. What we have to do is to consider the kind of system that is introduced when we put atoms into our empty box.

Sodium seems a good example to take, chiefly because its energy levels are well known from spectroscopic and other data for the free atoms in sodium vapour, and have survived the necessary forty years' probation that data must serve before getting into the better textbooks. I suppose I might also have chosen one of the other alkalis; or even mercury, though you would prefer a solid I think. Sodium, atomic number 11, has eleven electrons, two in the K shell, eight in the L shell, and the one that we shall be interested in, the valency electron, in the M shell. Fig. 1 shows many of the energy levels available to this No. 11 electron, the figures up the left-hand side indicating the total energy. This is made up of a kinetic-energy term which is always positive and decreases as we move upwards, and a potential-energy term which is always negative and also decreases numerically as we move upwards if we take the line that the arbitrary zero of potential energy is when the electron is at an infinite distance away. But as we move upwards, the total energy is increasing; although the +K.E. becomes less, the −P.E. becomes less negative, and it works out that the sum of the two becomes less negative.

So it is convenient to take the arbitrary zero of the *potential energy* to be such that, in the unexcited atom when the M electron is in its highest possible energy state, the *total energy* is reckoned as zero. This is fair enough, besides being a lot simpler, since only energy changes are going to come in. And, if we do have to consider any of the other ten electrons, we shall regard

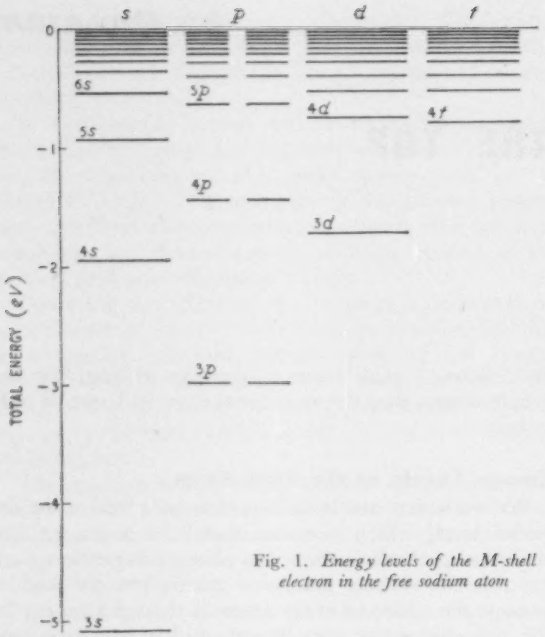


Fig. 1. *Energy levels of the M-shell electron in the free sodium atom*

this as being the reference zero of energy for the atom as a whole.

In Fig. 1, the principal quantum number is indicated by an arabic numeral; the azimuthal quantum number l by the letters s, p, d, f, for 0, 1, 2, 3, respectively; and the different azimuthal quantum numbers are tabulated in separate vertical columns—partly because this is the way in which the diagram has appeared for nearly half a century and to aid recognition, and partly because there is the point that energy changes take place only between adjacent columns. The p column, for the doublet level, appears in duplicate though the vertical separation is too small to show; for this indicates the start, within the free atom, of the kind of thing that happens with all the levels when we have large numbers of atoms very close together.

These energy levels represent eigenvalues of the wave-function, stationary non-radiating orbits, allowed energies, and so on, according to the language you are using. The spaces between them represent a continuum of non-solutions to the wave-equation, impossible orbits, or forbidden energies. While we are still discussing Fig. 1, we might even introduce the term 'band'; for all the white parts of the diagram represent forbidden bands.

In the splitting of the doublet p-levels, we get a foretaste of what is to come; they are slightly separated because of the effect of the rest of the atom on electrons of opposite spin. There is also some splitting (hyperfine structure) due to the magnetic effect of the nucleus. Further level-splitting (see "Photon-Spin Pumping", November 1959, p. 407) occurs in an external magnetic field (spectroscopically speaking, the Zeeman effect) and in an electric field (the Stark effect). The energy levels separate into bands, the structure and multiplicity of which are in this case determinable, and the region of forbidden states is by that much reduced. We can then forecast that the same kind of thing will happen to the levels of each atom of sodium when it is within the solid metal, for there will be a multiplicity of fields due to all the other atoms; and can expect that the energy levels will broaden into bands, each of them being not a continuum but a range of levels with very small but finite separation (Fig. 2). The Exclusion Principle shows us how the levels will split, since as they are to belong to the solid *as a whole*, and no two are to be identical, for N atoms there will be N different versions of the individual level.

This is one approach to the idea of band structure; it is clearly very much over-simplified, and must be revised in a moment. In particular, it deals only with the M-shell ('valency') electron, and there are all the others to consider. But we can see that the ground-state level of the valency electron may be replaced by a valency band—the inflated $3s$ level; and that above this come the bands corresponding to the lower excited states, the conduction band. At least, that is what we should *expect* and that is what some of the simple expositions would lead us to expect. Actually, and nobody ever seems to make this clear, when the atoms in the solid are close enough together there is, in the first place, a good deal of overlapping of the bands, and also a great convergence on the zero-line. The actual pattern for a given metal depends on the kind of atom and the spacing. For sodium, which here seems rather a bad example to choose for a general discussion, at the normal atomic-spacing distance of 3·67 angstroms, the $3s$ and $3p$ bands overlap and extend from zero downwards, $3s$ to about -8 eV and $3p$ to about -4 eV (Fig. 3). The higher ones seem to have disappeared into the region of positive energy which means that if there were any electrons within them they would be *free from the metal as a whole*, not just free within the metal. So sodium, in fact, gives what is effectively one large allowed band, in two pieces. Though Fig. 2 would be quite relevant at larger atomic spacings, of about 10 angstroms or so.

The Fermi-gas of electrons fits somewhere in the region indicated. There must be potential energy, or it would not stop there; if we imagine this provided by the potential walls of its box, the atoms themselves are responsible (now we have put them back) for furnishing

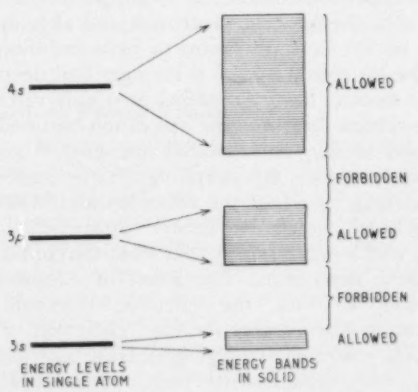


Fig. 2. *Development of bands as the Exclusion Principle operates for an assembly of atoms held together in a solid. In sodium, the atoms are actually so close together that the allowed bands overlap*

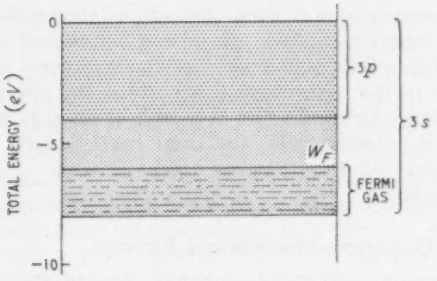


Fig. 3. *The only bands in solid sodium, with the approximate location of the Fermi gas and* W_F

these walls. And, now that we have the electrons in this condition, we ought to speak of the residue of the atoms as ions, which is what they are. It may seem odd to say that the energy is less than zero, when it is 'all kinetic'; but this condition always held inside the box, though we never mentioned the negative potential energy due to being in the box itself, because we never needed to when we were only considering the distribution of the kinetic energy.

The L-shell levels, 2*s*, 2*p*, and so on, are not appreciably widened into bands, even at this spacing; they lie at about 30 eV or so below the zero line, a quantum jump that corresponds to soft X-rays in the range that can be investigated with ruled gratings. And this means that it is possible to see what is really happening to the Fermi-gas in sodium.

Finding the Fermi Level

In an article on "The Soft X-ray Spectroscopy of the Solid State", published by the late H. W. B. Skinner in the 1939 volume of "Reports on Progress in Physics", there is an account of experiments which locate the Fermi level, and also confirm the shape of the Fermi distribution curve. These work in both directions, absorption and emission, and results are given for the sodium Na L radiation.

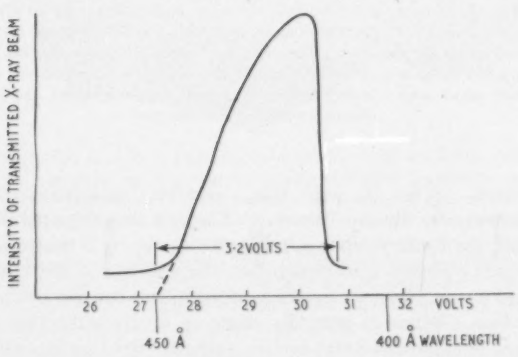


Fig. 4. *Falling off of absorption, evidenced by increased transmission, produced by occupation of levels in the 3s band by Fermi-gas electrons. The 'volts' represent the exciting voltage of the X-rays. The trouble extends over a range of about 3.2 volts*

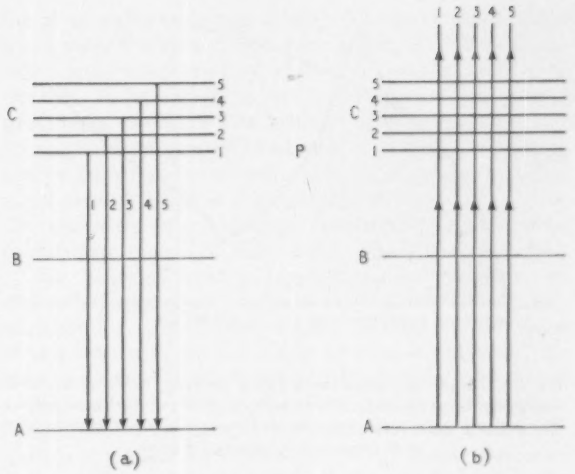


Fig. 5. *Illustrating how Fig. 4 happens; (a), the emission of lines in the L-series X-ray band by the return of electrons excited into the C band. The reverse process of absorption would be expected. But if X-rays of exactly these wavelengths ejected electrons from the L-shell level A, they would have to lodge in exactly the same levels in the C-band. They cannot do this, as the levels are occupied by the Fermi gas. So the process cannot happen, and the X-rays are not absorbed (b)*

The absorption of an X-ray photon by an electron in the L shell must, in order to happen at all, raise the electron either to a higher energy level in the solid or else to a sufficiently high energy to eject it completely.

Fig. 4 shows the way in which the intensity of the radiation transmitted through a sample of sodium depends on the wavelength (and on the original exciting voltage). Just below 28 volts, the transmission begins to increase, for electrons of this energy are beginning to find the level they seek occupied by members of the Fermi gas (Fig. 5). This goes on, getting worse and worse, until a little under 31 volts. Then over the precipice of the Fermi level and back to normal again. The width of the hump is very nearly 3·2 volts. The volts represent total energies measured with *respect to the L levels*. Fermi-gas kinetic energy has its zero at about the point P. So that, from this point of view, the L levels have a total energy of something like −28 volts.

Skinner's article also gives the intensity contour of the L X-ray emission for sodium, which is very similar in shape to the hump on Fig. 4. Here, after electrons have been ejected from the L shell by electron bombardment, the vacancy is filled by one of the Fermi-gas electrons from the conduction band falling into it. The probability of this happening for a particular energy depends on the density of electrons in states close to this energy, so that once again the distribution of intensity follows very closely that of the Fermi distribution curve, and has a width of about 3·2 volts.

How closely experiment at ordinary temperatures fits in with the Fermi-gas for absolute zero can be seen from Fig. 6 where the Fermi-distribution curve is shown as a dotted line. This is really Fig. 3 turned through a right-angle, with the distribution contour superposed.

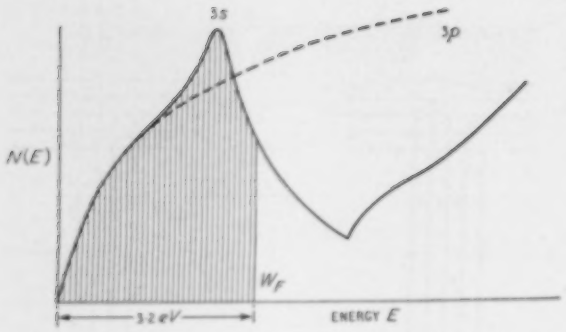


Fig. 6. *The full line shows the density of the energy levels, N(E), in the overlapping 3s and 3p bands. The dotted line is the Fermi distribution function. The vertically shaded region represents the Fermi gas. Turned round and shorn of its curves this is really Fig. 3 again*

The Mechanism of Conduction

We need not, for the moment, know how many of the conduction electrons comprising the Fermi gas have conduction-band energies, though the distribution of Fig. 6 has, of course, been worked out. All that concerns us is that there are finite energy steps possible within the Fermi distribution, and finite energy steps within the conduction band which will be filled by the Fermi-gas electrons as far as the two regimes overlap. An electron receiving a quantum of energy will be raised to a level above W_F; and within the conduction band there is a vacant place left for it to fill (Fig. 7).

It seems hard at first to grasp that conduction is a quantum effect. But if you apply an electric field, the electron moves in the appropriate direction, acquiring kinetic energy; it can only do this in quanta; which means that it is raised to a higher energy state and, for thi to happen, the state must be there available for it to enter. The lattice ions come into the dissipation of this energy, which is converted largely into thermal-agitation energy, so that there is a sort of detailed balancing, with as many electrons moving up from a lower to a higher state as return from the higher to the lower state; remembering that all the time they have the higher energy they are on the move and doing their job of conducting.

Conduction, in short, has the same general features as any other process for the absorption and re-emission of energy on the microscopic scale. It can be likened to the excitation of a resonance line, except that in that case, for radiation of frequency ν, the upper level must be exactly $h\nu$ above the lower one every time and the excitation is by radiation (though, of course, the early experiments of Franck and Hertz did exploit inelastic collisions). The reason why you can only get it with free atoms in the gaseous state is because this sharpness is necessary. In conduction in a solid, the energy is drawn from the electric field, and there must be *some* level of the appropriate energy to accept it while it does its conducting.

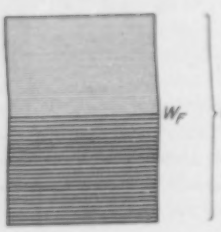


Fig. 7. *The condition for conduction; vacant allowed energy levels above W_F*

The Quantum-Mechanical Picture

There is another way in which we can arrive at the band-structure of a metal. This is to take the wave-function for *every* electron as extending over the whole region of the solid, and that for *every* nucleus as well, and to sort out the energy ranges within which possible solutions for the wave-equation for an electron are to be found. (Or, if you like, possible stationary-wave patterns for electron-waves.) This is a lengthy operation and, for some purposes, has to be done; it is found that bands overlap, and depend on the spacing of the atoms, and so on. But, at least for sodium, less drastic methods suffice. What is actually done is to consider the *conduction* electrons only, in the potential field produced by the *positive ions* [Fig. 8 (*b*)]. This is all that is needed to furnish the required information. And in any case, the inner levels of the atoms are far below the ground-state of the free atom, and there is no point for this

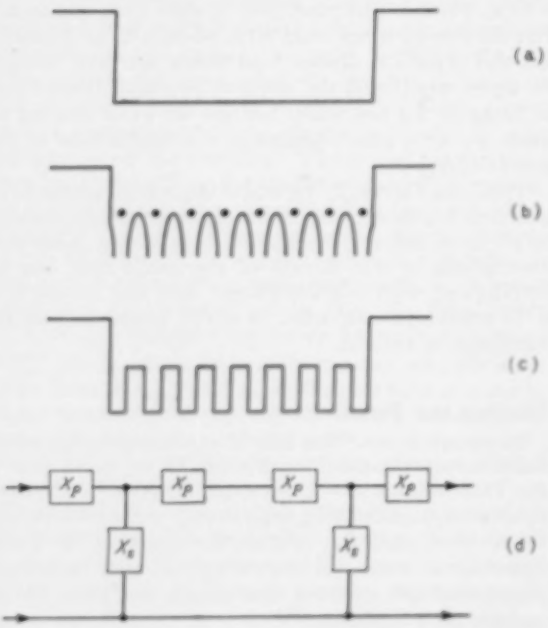


Fig. 8. (*a*) *The potential well of the original Fermi gas of 'electrons in a box' at absolute zero;* (*b*) *the periodic potential variation throughout the crystal lattice, shown in one dimension only, due to the positive ions;* (*c*) *the further simplification of* (*b*) *for mathematical treatment, when it turns out that its action on the electron waves (this term is used rather than 'wave functions') is analogous to that of a periodic filter ladder of reactances or a transmission line, which passes some wavelength bands (allowed electron energies) and stops others (forbidden electron energies)*

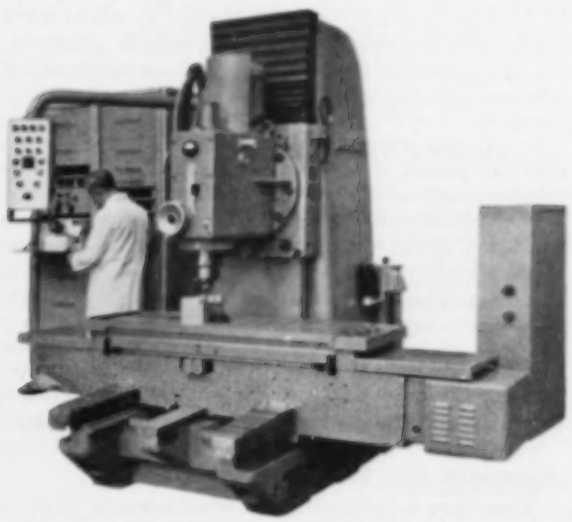
purpose in trying to calculate imaginable movements of a particle as effectively immobilized as this. "Every electron" means all the electrons that are contributed to the Fermi gas.

The treatment of the case of Fig. 8 (c) is outlined in Bleaney and Bleaney, pp. 522 sqq., after first rounding off the hops of Fig. 8 (b) to make a rectangular periodic potential pattern. This is a distribution in one direction only, and it must be remembered that even this has to be extended to three dimensions for an actual metal if it is to mean anything. But the way in which the wave-function behaves can be visualized without any further calculation, for the periodic structure of Fig. 8 (c) can be likened to the ladder network of a filter (Fig. 8 (d) which will transmit some wavelengths and stop others. For its effect is to allow electrons of certain wavelength (and hence energy) ranges to exist in the solid, in the allowed bands, and to stop other ranges, which form the forbidden bands.

To sum up, whether you build up your crystal of metal an atom at a time, so that the energy levels of the free atoms are spread out into bands, or whether you take a solid lump of metal at the start and consider each electron as belonging to the metal as a whole which provides a potential box with a periodic pattern, the arrangement of allowed (and also, in general, forbidden) energy bands presents itself. Motion of electrons under small electric fields is only possible if they can acquire the necessary small energy increments by stepping upwards in a partially filled band. And the distribution of the energies among the conduction electrons at ordinary temperatures follows very nearly the Fermi-distribution. And I think I have an analogy, for those of you who may be travelling (if that is the term) by road to the seaside this Whitsun. If you think of yourself at the head of the queue at a side-road as representing a Fermi-level electron, and the main road as a periodic pattern of allowed and forbidden energy bands, you may eventually get yourself conducted if you achieve just the right kinetic energy to land in the middle of an allowed band.

ELECTRONICALLY-CONTROLLED MACHINE TOOL WITH ERROR DETECTION

The illustration shows a vertical milling machine fitted with the latest E.M.I. numerical control system incorporating a unit for the detection of errors which have occurred during the production of the punched-paper control tape.



With this system, a punched-tape reader feeds co-ordinate data to an interpolating unit which, in turn, provides continuous positional control signals to the machine axes-control motors via an intermediate store. The angle-error unit monitors the positional information in the intermediate store and, if the information indicates that the course of the cutting tool is about to deviate outside a preset angle limit, the machine is switched off before the cut is taken. It is arranged for the angle-error unit to be inoperative at any point in a programme where a definite change of direction of cut is required.

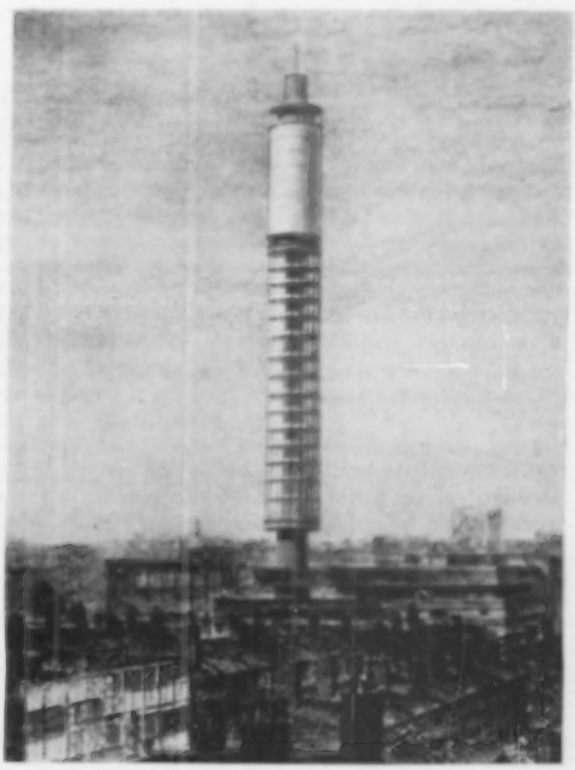
ELECTION TO THE ROYAL SOCIETY

Professor H. E. M. Barlow, Dean of the Faculty of Engineering, Pender Professor of Electrical Engineering, and Director of Laboratories, University College, University of London, has been elected a Fellow of the Royal Society.

500-FT. RADIO TOWER FOR LONDON

This artist's impression shows the 500-ft. tower for television and radio-telephony which the Post Office proposes to build at the Museum Telephone Exchange, London.

The new tower will not be used to broadcast television programmes direct to the public. It will be used to relay programmes between the studios, control centres and transmitters of the broadcasting authorities. A public observation platform is to be provided at a height of about 460 ft.



New Distortion Criterion

PART 1: AUDIENCE REACTION TO NON-LINEARITY DISTORTION

By E. R. Wigan, B.Sc.(Eng.), A.M.I.E.E.*

SUMMARY. *In any sound-transmission system non-linearity distortion is always present to some extent. The optimal design of line amplifiers, amplifiers for loudspeakers, modulators for radio transmitters, and the like, cannot be achieved until the rate of exchange has been established between the degree of distortion and the resulting deterioration of programme quality. In this article an attempt is made to establish this rate of exchange.*

Part 1 analyses the subjective aspect of the problem: experiments are described which separate out those features of the distorted sounds which are responsible for the audience reporting that they are unpleasant to listen to. Almost without exception the data conform to the comprehensive statement that the unpleasantness of a distorted sound is judged in terms of 'the time-rate-of-departure of the signal from normality'. It is found that listeners are influenced by pitch, by loudness, and by contrast effects, and that they have very short-term memories of distortion phenomena.

Distortion is commonly assumed to be important only to those concerned with high-quality sound reproduction; it is almost entirely overlooked that it has a direct influence upon apparatus design. For example, the power-handling capacity of any piece of equipment dealing with programme signals is ultimately settled by the relationship between power and distortion. The designer has to know how much distortion can be tolerated, for the most efficient design will be the one which will be just on the verge of exceeding this tolerance-limit when on full load.

Failure to achieve this state of affairs is serious in cases where the cost of the equipment is high in the first instance, and rises rapidly as the power-handling capacity is increased. A typical example of this is a high-power modulator for an a.m. radio transmitter and, in passing, it is to be noted that such transmitters are not only used for the medium-wave service but are in demand for the sound channels of television transmitters. The designer's objective is clear enough, and the distortion produced at full load is readily calculable; but no satisfactory method has been found which will convert the distortion percentage into units of *subjective* distortion or, what is called in this paper, 'unpleasantness'.

Designers are well aware that conventional distortion measurements do not allow the unpleasantness to be predicted, except in very general terms and, for lack of a more reliable quantity, reinforce their specifications in other ways. For example, the usual form of specification to which B.B.C. radio transmitters are designed will stipulate the maximum tolerable total harmonic distortion at various modulation-depths but, at the same time, it will demand that the transmission shall *sound* satisfactory. That the harmonic distortion percentage does not correlate with the unpleasantness is shown also by two transmitters which yield virtually the same percentage total harmonic distortion, but one may give just acceptable and the other quite outstanding performance. Clearly a criterion is required, which may be quoted in a specification without need for reinforcement. To be of any value however, it must be shown to agree closely with the opinions given by reliable observers. The first move therefore is to discover those factors which influence the listeners when making distortion judgments.

In this part of the paper it will be demonstrated that although skilled observers are consistent in the sense that they respond in the same, or nearly the same way, to the *same* sounds, their reactions to distorted sounds which are unlike in tone-colour, pitch, or loudness are in some respects quite unexpected. The elucidation of these reactions made possible the formulation of the distortion-criterion which will be discussed in Part 2 for, besides having the duty of translating into numerical terms the subjective 'rating' of the objective distortion, that criterion must stand up to the test that it reflects to some extent the behaviour-patterns now to be illustrated.

Apparatus and Test Techniques

Tests were conducted in an office roughly 16 ft. × 18 ft. × 9 ft. converted into a listening room by adding acoustic damping so as to be roughly representative of a sparsely-furnished living room. The noise level was checked periodically and varied between about 30 and 40 phons. It was decided that the sound-reproducing equipment should be of the highest quality, so that the crew of experienced observers would be in a position to make distortion judgments which were both severe and repeatable.

It was realized that it would be very difficult to train observers to rate the degree of distortion in terms of some scale, and so the experiments were designed from the outset so that only 'equality' judgments were asked

* B.B.C. Research Department.

for. In this way the observers' skill in distinguishing very small differences of distortion was exploited, while their personal opinion as to the *degree* of difference carried no weight.

A diode was used as the distorting medium throughout the investigation which, in this respect, differs radically from others previously carried out[1,2]. It might at first appear quite unrealistic to use distortion of such a gross kind but, as explained in Appendix 1, the distortion generator is designed so that the diode distortion can be 'diluted' down if need be and minor distortion conditions can then be represented. Moreover, if the behaviour of any conventional high-level modulator for an a.m. radio transmitter is examined it will be found that the distortion, although very considerably attenuated by negative feedback, is of typical diode type, arising generally from grid current. In the present tests the 'dilution' operation takes the place of the negative feedback.

There were however several unpremeditated advantages in using a diode; in particular, it was found possible so to arrange the distortion generator that when programme signals of continually fluctuating amplitude were being distorted *the distortion percentage was independent of the signal amplitude* [see App. 1 (c)]. This very unusual state of affairs made possible the work described in a later section ("The Audience Reaction to Distorted Programme Material").

Two classes of test signal were adopted, one was a short section of the programme signal itself, and the other was a pulse of pure tone; (see Fig. 1). This was chosen for two reasons; because it gave to the ear the same general impression as a staccato piano note; and also because the circuit which produces it (described in App. 2) has the advantage that, by switching off the battery, a steady signal of precisely the same amplitude as at the crest of the pulse is generated. On this steady signal distortion measurements can be made without difficulty.

The corresponding sound level was generally adjusted to be 70 phons during the tests. The circuit has the further advantage that the width of the pulse T_p

LIST OF SYMBOLS

S	=	The step number to which the distortion-selecting switch is set. ($S = 10$ implies that the diode distortion is presented completely undiluted.)
S_A; S_B	=	Values of S appropriate to the 'A' and 'B' test signals.
S_{BM}	=	The median value of S_B derived from the intersection of the scoring function with the 50-percentile line.
V_b	=	The bias voltage applied to the distortion diode, and which must be exceeded by the input signal before distortion begins.
a.d.r.	=	The 'apparent distortion ratio'. The a.d.r. of the 'B' relative to the 'A' signal is S_A/S_{BM}; and the inverse of this is the a.d.r. of the 'A' relative to the 'B'.
T_p	=	The width in seconds of the test pulse measured at half its height.
G	=	The grading (on an arbitrary subjective scale of 0–5) of the distortion present in a specific 5 seconds of a programme item. (G is the average of the gradings assigned by a crew of observers.)
G_0	=	The value of G calculated from a group of items which have in common the value of S. (G_0 is taken from a best-fitting line.)
C_t	=	The distortion criterion calculated on the basis that no harmonic is audible unless its percentage exceeds t, and that up to ten terms are summed.
C_0	=	The criterion calculated on the basis that t is zero, and 10 terms of the harmonic series are summed.
$(C_t)_A$; $(C_t)_B$	=	The values of C_t appropriate to the A and B signals.
n	=	The harmonic number ($n = 1$ implies the fundamental).
p_n	=	The percentage ratio of the amplitude of the nth harmonic to the amplitude of the originating sine wave.
t	=	Ideally a function of frequency but, in this article, a fixed quantity: t is the quantity to be subtracted from the (un-weighted) amplitude of each spectral term, to allow for 'masking' and noise, before C_t can be computed.
A	=	Amplitude of the basic sine-wave signal.
T	=	Period of signal wave.
d	=	That part of the period during which the distorting diode is effective.
q	=	That fraction of 'A' over which the distorting diode operates.

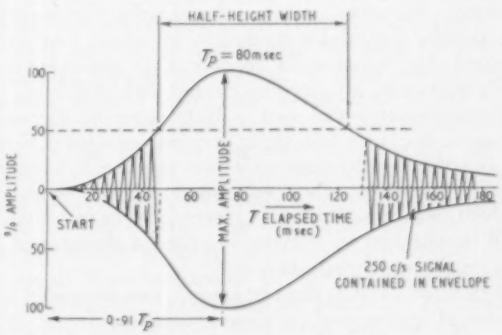


Fig. 1. *Prototype pulse envelope. The proportions of the pulse envelope in terms of T_p, the half-height width, are independent of T_p. Width can be varied from* 10 *msec upwards by the generator of App.* 1

measured at half-height, can be varied continuously from 10 to 65 msec; generally it was set at 40 msec.

It is important to note that in all the experiments the pulse was generated first and was distorted later (see App. 1) so that, if the distorting diode was heavily biased, only the upper part of the pulse was distorted and both the initial and final sections of the pulse remained quite unaffected.

Arrangement of the Pulse Experiments

The high-grade loudspeaker used throughout the investigation was of the type employed for quality monitoring within the B.B.C. Hidden behind a thin curtain, lit from above to make it invisible, the loudspeaker stood in a corner of the room with the crew of six observers in two rows about 6 ft away. Inevitably, a wave-interference pattern was generated in the room, and so some observers heard the sound as having a larger or smaller fundamental component and, consequently, reported it as less or more distorted than other

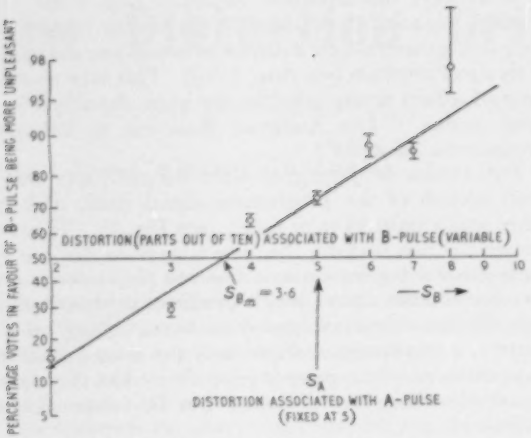


Fig. 2. Typical scoring function. The distortion S_B in the B-signal is varied. Six observers report whether B is more or less unpleasant than signal A. The mean line shows (at S_{BM}) where equality is reached. Bars thus: Φ are explained in App. 3 (d)

observers. However, because the experiments involved two sounds, the *relative* distortion of which had to be observed, and because the reports of six observers were finally averaged, interference caused little difficulty. In most experiments the two sounds were of the same frequency but, in a few, their frequencies were different and the above argument no longer applied to the same extent. However, the averaging of the observers' opinions (see App. 3) led to a representative result which could be checked by moving the crew a distance of $\lambda/2$ to or from the loudspeaker.

In general, each of the test sounds was presented with a pulse of undistorted (or clean) sound before and after, so that the aural context of the distorted sounds was well defined. The test sequence was always duplicated so that the observer could, on the second hearing, check the opinion he had reached on the first; it ran:

CC – AA – CC – – CC – BB – CC – – – CC – AA –
CC – – CC – BB – CC

where 'C' indicates a clean pulse of sound, and 'A' and 'B' the two distorted versions of the pulse. The audience had to say which sounded the more unpleasant. The dashes indicate equal time intervals of between 0·3 and 0·4 sec. The sequence above having been played through, the distortion of the B signal was changed and the sequence repeated, and so on.

It was soon noticed that when two sounds were nearly equally unpleasant the observers had great difficulty in reaching a decision; especially if there was a marked difference of tone-colour. However, by introducing the category rating mentioned in App. 3 it was possible to produce a scoring function such as Fig. 2; this gave a very good idea of where the equality point lay, in spite of the observers' reluctance to give equality judgments.

The observers reported that their attention was waning at the end of a single-test series which, to provide six test points as in Fig. 2, would require twenty-four sound comparisons, and was very tiring. However, the introduction of the apparatus described in App. 3, which registered the audience response anonymously, made things easier; for listeners give their opinions much more freely when no record of individual response (which they fear might reveal gross inconsistencies) is being made. At the same time the apparatus eliminated the clerical errors associated with the marking of score sheets and the abstraction of the data.

Method of Varying the Distortion and the 'Difference Signal'

The distortion produced by the generator described in App. 1 can be varied in any desired proportion by 'diluting' the characteristic diode distortion with undistorted signal. In Fig. 2, the quantities S_A and S_B, on the X-axis express (in parts out of ten) what fraction of distortion is present in the A and B signals, S being variable in unit-steps from zero to 10, at which point the distortion is entirely that generated by the diode.

Fig. 3 illustrates the change of wave shape as S decreases from its maximum (i.e., dilution increases from zero), the diode being biased by a voltage V_b. Notice that the dilution does not extend to the bias voltage V_b but only to that section of the wave which exceeds V_b. The difference between the distorted signal and the undistorted signal is called here the 'Difference-

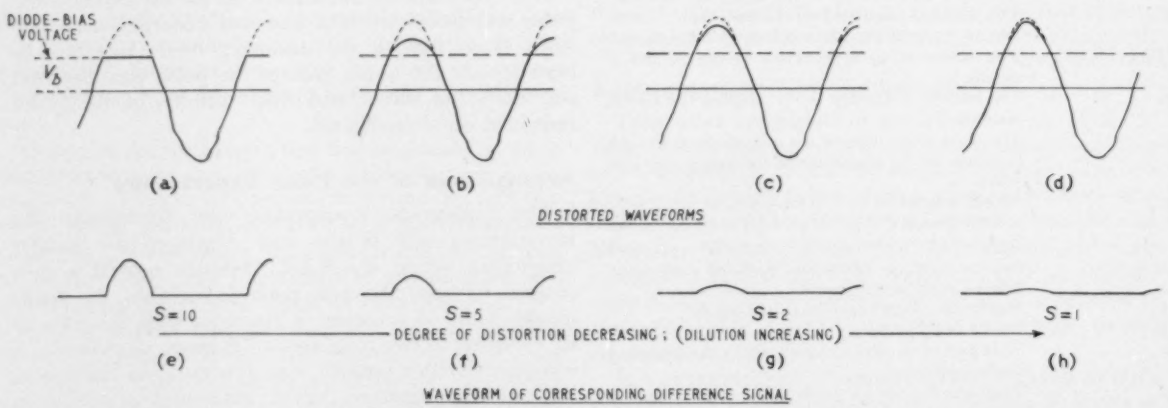


Fig. 3. The 'dilution' of distortion and the 'difference-signal'. Changes of wave-form as diode distortion (a) is diluted by adding undistorted signal. Subtracting the full-line curve from the dotted one gives the 'difference-signal' representing the distortion present

Signal' and is shown at Figs 3 (e) to (h). Difference-signal amplitude is strictly proportional to the value of S, other things being equal. Dilution causes the amplitude of every term of the distortion spectrum to shrink equally and, consequently, there is no fundamental change of tone colour—a most important point; it helps the listener to make accurate equality judgments.

The Apparent-Distortion Ratio

Fig. 2 illustrates a purely imaginary experiment and from it it can be deduced that the B signal was more unpleasant that the A signal, for the diagram demonstrates that the B signal, when distorted to the extent indicated by $S_{BM} = 3{\cdot}6$, sounded exactly as unpleasant* as the fixed A signal, the A signal itself being distorted to the extent $S_A = 5{\cdot}0$; thus it needed only 3·6 parts of the B-signal distortion to equal 5·0 parts of the A distortion. In a certain sense, therefore, the undiluted B signal could be stated to be *intrinsically* more unpleasant than the undiluted A in the ratio 5·0/3·6. This ratio is called here the apparent distortion ratio (a.d.r.) of the B relative to the A signal. For lack of any reliable *subjective* measure of unpleasantness we shall refer to the a.d.r. in speculating about the significance of later experiments, but it must be borne in mind that it is derived from a purely *objective* measure—the ratio S_A/S_{BM} of degrees of non-linearity.

The bars shown in Fig. 2 above and below the plotted points relate indirectly to the precision of the data. They are explained fully in App. 3 (d).

Experiments Leading up to the Deviation-Rate Hypothesis

Examination of the Difference Signal

When a sine wave is 'clipped' by a diode the various waveforms shown in Fig. 4 (a_1) to (d_1) are generated as clipping is increased. Corresponding difference signals appear at (a_2) to (d_2). In an extensive series of subjective tests each of these distorted signals was compared with signal (a_1), which was first diluted by 30% [see (a_3)]. It was then found that signals (b_1), (c_1) and (d_1) had also to be diluted to various extents, as shown at (b_3), (c_3) and (d_3) in the third line of the figure, in order to make them as unpleasant as the first signal (a_3). The fourth line shows the difference signals which correspond to the resulting four equally-unpleasant sounds†.

Two points at once emerge:

(1) the areas under the difference signals, (a_4) to (d_4), are not equal—*implying that the unpleasantness* is certainly not related to the power level of the *distortion terms; notice also that, as a rider to* this, the total harmonic percentage cannot be related to the unpleasantness; *and*

(2) the slopes of the difference signals (a_4) to (d_4) are all very nearly the same, as indicated graphically in the last line of Fig. 4, and in a more precise way by the straight line in (f), derived from the parameters defined in (e).

Now the amplitude/time function of the sound-pressure wave which reached the listeners in these experiments certainly bore no simple relationship to the voltage/time picture set out in the figure—phase distortion both in the generation and the transmission of the sound must have distorted it grossly—nevertheless the diagram suggests that the unpleasantness is related in some way to the time-rate-of-change of the difference signal. In Part 2 and App. 5, this anomaly is resolved.

The Deviation Rate Hypothesis

Repetitions of the last experiment at a variety of signal frequencies and with different values of the bias voltage V_b and of the pulse width T_p, and using several different crews of observers, all led to the same general conclusion, and from this point the 'Deviation Rate Hypothesis' began to take shape; it ran as follows:

> 'The unpleasantness of a distorted signal is assessed in terms of the time-rate-of-deviation of the signal from normality'.

If this hypothesis is sound then, in experiments in which the observer hears the 'clean' in addition to the distorted version of the signal, the term 'normality' should be interpreted as the original *sine-wave from which the distorted signal was derived*, and it has to be postulated that the listener takes the amplitude of this wave as his basis of reference in weighing up the distortion he hears. Experimental evidence which seems to support this postulate is set out in the App. 5. Again if the listener has no immediately-available standard sound to listen to he must necessarily make a guess at what 'normality' is. There can be no doubt that it is on this *guess* that he bases his opinions of broadcasts, gramophone recordings and the like.

A Crucial Experiment

When the signal was distorted by *two* diodes, one acting upon each half of the wave, and particularly when the diodes were set to clip the two halves symmetrically, the result was quite unexpected—the unpleasantness *was the same as with one diode alone*. Moreover, when the clipping was asymmetric the smaller degree of clipping contributed nothing to the unpleasantness although there was, of course, in this and the other case, a gross change in tone colour.

For a long time this evidence held up further progress, for it seemed to discredit completely the deviation-rate hypothesis, no matter how it might be re-phrased. For instance, it was argued that the appearance of distortion twice per cycle of the wave would imply a doubling of the 'rate of occurrence' of distortion which, from the hypothesis, should have made the distortion more unpleasant. The explanation did not appear till much later—Part 2—but the search for it led to the most careful re-examination of cases such as Fig. 4, and the establishment of that phenomenon beyond any doubt. At the same time the hypothesis was reinforced by the experiments now to be described.

* The 50%-score (Y-axis) in favour of 'B' as the more unpleasant implies an equal score for 'A'; ergo 'B' is as unpleasant as 'A'.

† Here is should be remarked that, so far as could be ascertained, judgments of the unpleasantness of distorted sounds, depend upon the *maximum* value of the distortion within the signal cycle; this coincides with the crest of the test pulse.

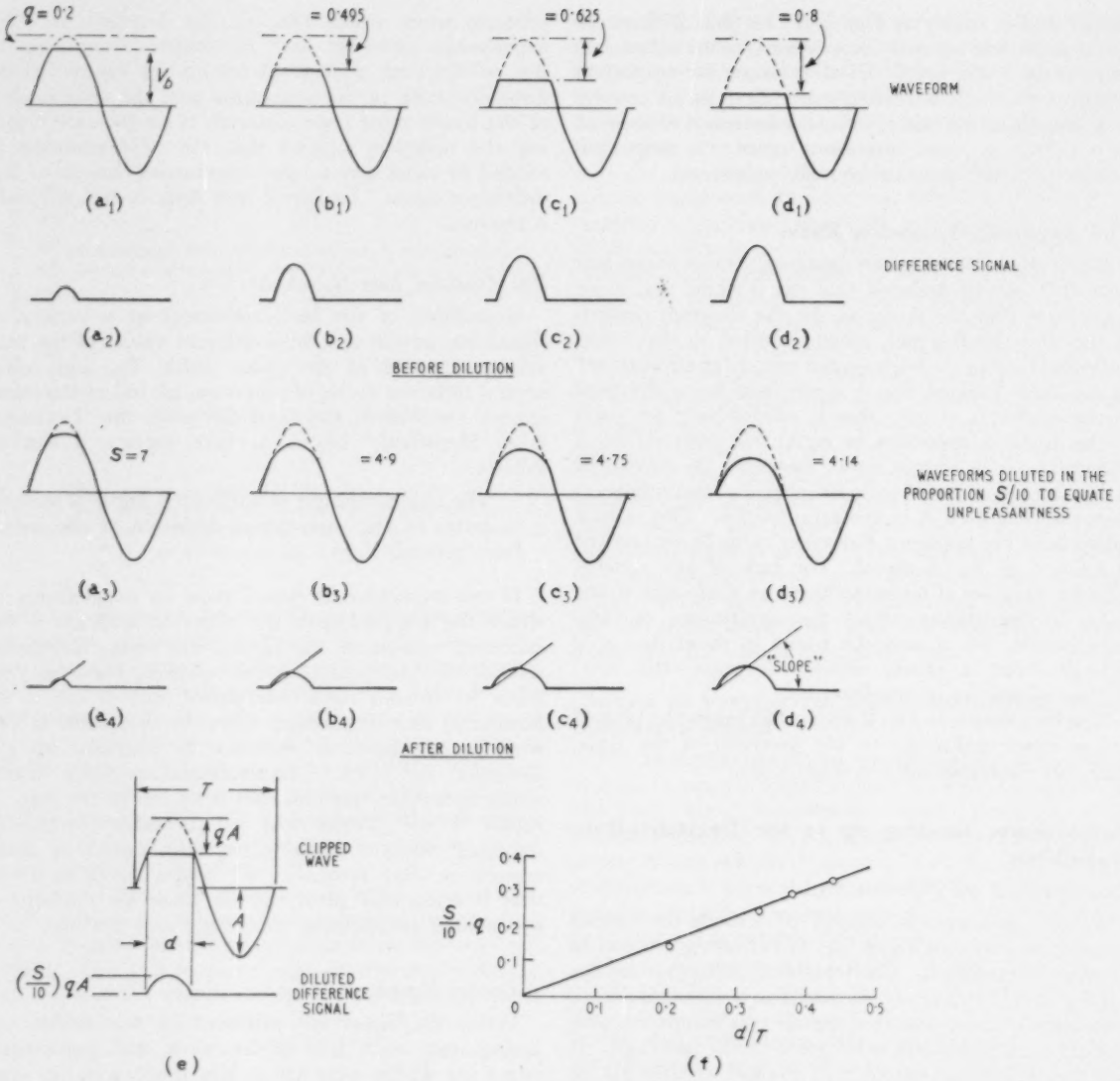


Fig. 4. *'Clipped' sine waves which sound equally unpleasant. Altering the diode-bias V_b changes the degree of clipping; altering the dilution equalizes the unpleasantness of the four signals. At (f) the slopes of the difference signals a_4–d_4 are shown to be nearly equal. Terms defined at (e)*

Experiments Suggested by the Hypothesis

The first experiment the outcome of which could be logically deduced from the hypothesis was as follows:

That the unpleasantness is a function of the slope of the difference signal implies that the unpleasantness of a given wave-shape would be increased or decreased by a corresponding change in the frequency of the test tone. To test this, two signal pulses, equally distorted, and of the same amplitude, but differing by an octave in pitch (500–1,000 c/s) were compared; as expected, the higher-frequency pulse was the more objectionable. On further examination the effect was found only when the octave interval lay in the range 250/500 c/s to about 1,500/3,000 c/s. This frequency-effect was most strongly marked when the octave was 500/1,000 c/s, where the distortion of the higher-pitched pulse had to be diluted in the ratio of about 3·2/1 (see Fig. 5), compared with the lower-pitched to make the pulses equally unpleasant.

Such tests were very difficult to carry out with precision, for irregularities in the loudspeaker response influenced the results; the curve in Fig. 5 is a kind of average obtained with a number of loudspeakers. For these reasons, too much weight must not be placed upon the values of a.d.r. given in the figure, but the disappearance of the phenomenon in the 1,500-c/s region is clearly demonstrated. It seems likely to be no accident that this corresponds to the region in which there is a radical change in the transmission characteristics of the aural nervous system. Although

the low-frequency limit to the phenomenon (at 250 c/s) does not coincide with any corresponding change of nervous response, it is confirmed indirectly by data presented in Ref. 2.

A second, and very simple way of increasing the slope of the difference signal would be to increase the *amplitude* of the distorted signal; that is to say, the sound has merely to be made *louder*. A direct test for the existence of this effect was completely unsuccessful and for an interesting reason. On discussing the results with the observers, it was found that they were unable to compare the louder with the softer sounds as instructed; instead, they found themselves comparing each of the test sounds with an *imagined* clean signal of *equal* loudness. Later (see Section below) the 'amplitude effect' was clearly demonstrated under less academic conditions, when the pulses were grouped into short melodic phrases, and when one of them was made louder than the others.

The proposition that unpleasantness was judged as *deviation from normality* was confirmed in the course of an experiment as follows: the test-signals were not single pure tones but a mixture of two 'locked' pure tones, the second being the octave of the first and of half the amplitude. This complex wave was asymmetric and produced difference-signals of radically different slopes when distorted by a diode, depending on the 'sense' of its connection. The experiment confirmed that the more steeply-sloped curve was the more unpleasant. But, afterwards, the observers remarked that they were almost unconscious of the 50% second-harmonic in the 'clean' signal once they had heard it in comparison with the distorted version The subjective data confirmed this by revealing a consistent audience response to quite tiny degrees of distortion of the complex waveform.

Finally, the hypothesis suggested indirectly that the rate-of-rise of the pulse of tone itself might have a bearing upon the unpleasantness. This was not so, for careful tests showed that a change of pulse-width T_p from 40 to 75 msec. had no effect upon the unpleasantness.‡

‡ The result of this series of experiments must be treated with caution however, in view of the possibility that, as in the tests for the amplitude effect, the audience might have failed to carry out their instructions.

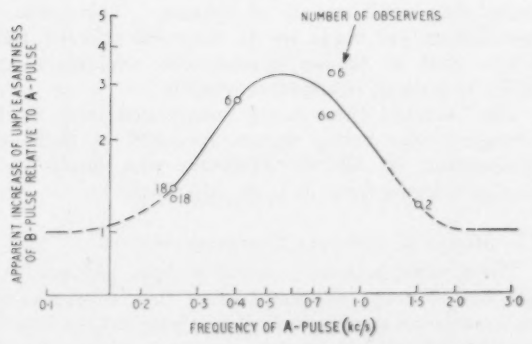


Fig. 5. *The frequency effect. Distorted pulses of tone sound more unpleasant if their pitch rises; the rise here is one octave*

Elementary Forms of Programme Material. The Influence of 'Context' upon the Unpleasantness

Test Material

An elementary electronic 'piano' was built, capable of ten notes of the C-major scale, and producing pulses of tone rather like those shown in Fig. 1. § No chords could be played, only phrases of single notes. By an artifice, the 'tail' of each pulse was removed by initiating a rapid acceleration of the decay of the signal directly the key was depressed for the following pulse. In this way, a fairly rapid rate of playing was feasible without introducing audible distortion of the pulse.

A complete phrase was presented under the test-condition 'A', followed by the same or a different phrase under condition 'B', and, as no undistorted versions were presented, all judgments were based entirely on *the observers' estimate of 'normality'*. In this respect, the experimental conditions began to approach realism.

'Contrast' Phenomena

Under these new conditions, the observers' judgments depended critically upon the 'contrast' between the sounds heard and the sounds expected. The expectation could be altered by altering the general level or background of distortion.

For instance, in the experiment illustrated in Fig. 6 melody A and melody B were presented in sequence; each took about 1·5 sec to play through, and there was a silent gap of roughly 0·5 sec between them. A background of distortion was set up using the initial notes;

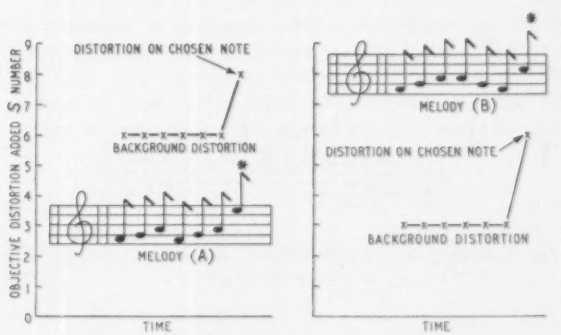


Fig. 6. *'Contrast' effect. Degree of distortion of the preliminary notes decides the unpleasantness of the final ones*

it was twice as big in A as in B (6 as against 3 units). The final note of each phrase was rather more distorted; in the case of A it was fixed at 8 units, and the distortion of the final note of B was adjusted during the experiment.

The observers were asked to say when this adjustment made the two final notes equally unpleasant. It was found that the observers had chosen 6 units in

§ The reason for the difference in pulse shape is explained in App. 2.

the final note of the B melody as sufficient to balance 8 units in the A melody.

It is clear that the listeners were influenced by the contrast between the distortion of the final notes as compared with the background of distortion in each phrase; six units of distortion were quite sufficient in the B melody (when contrasted with the background of 3-units) to sound as unpleasant as 8 units when contrasted with 6, in the A melody.

Test for Memory

The influence of the shortness of the observers' memory of distortion was traced in many experiments, and an attempt to establish the memory time-constant was made as follows.

The fifteen-note phrase of Fig. 7 was used. The audience was told to say when the two marked notes (of the same pitch G, to make them easily recognizable) sounded equally unpleasant. The distortion of the first of them was fixed, and that of the second was varied.

To make sure that casual variation of the ambient noise while this long phrase was being played would not vitiate the experiment, the intermediate notes were somewhat distorted (5 units) but less than the marked notes (7 units).

The audience asked that the final G should be distorted by only 6·25 units to make it sound equivalent to the first G (7 units); it was evident that the memory of the first hearing had faded in the interval (about 3·5 sec) before G was sounded the second time. However, the degree of fading is only slightly less than unity, $\frac{6{\cdot}25}{7{\cdot}00} = 0{\cdot}89$, which indicates a very large time-constant. Using the background-distortion level as datum the ratio becomes much smaller; i.e., $\frac{6{\cdot}25 - 5}{7 - 5} = 0{\cdot}625$; which suggests a time-constant nearer to 3·5 sec. On other occasions, and by various other means, time constants varying from 1 to 10 sec were observed.

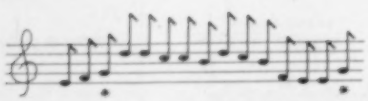


Fig. 7. *Memory test. The marked notes, though equally distorted, do not sound as if they were*

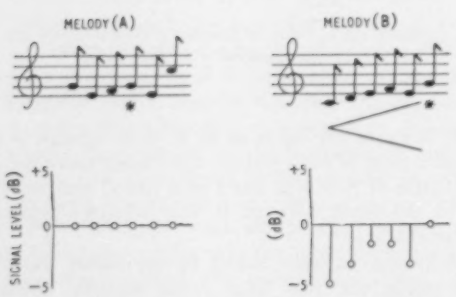


Fig. 8. *Amplitude effect. The crescendo makes the marked note more unpleasant*

Amplitude Effect

The amplitude-effect was clearly in evidence in the test material given in Fig. 8. The curve below the notes shows how the gain of the loudspeaker amplifier was varied in the second phrase so that it progressed with an interrupted crescendo terminating in a 4-dB lift on the last note, G. The G in the first phrase had no stress. The audience asked that the distortion of the G in the second phrase should be reduced to 70% of that in the first to make these notes equally unpleasant.

Notice that these two sounds were heard at precisely the same level, and had the same pitch; the distortion reduction called for by the audience is therefore almost entirely attributable to the increase in unpleasantness arising from the crescendo in the second case. Unfortunately, it was not appreciated at the time that a decrescendo embodied in a third phrase would have separated the amplitude effect from the influence of memory decay.

The Audience Reaction to Distorted Programme Material

Test Material and Conditions

Complete realism was impossible, for the programme items had necessarily to be short. The test items were selected by extracting 5-sec sections from a continuous tape-recorded programme, alternate 2- and 9-sec periods being blanked off. (The longer gap was to allow the observers to review their opinion of the previous pair.) Continuity was thus maintained; the listener knew roughly what to expect but the silent gap prevented him from carrying his impression over to the next item. Each item was distorted by a pre-set amount chosen from six or seven values in no regular order; completely undistorted items were included.

The observers were told this, but they were given no demonstration of such items nor were they told how many there would be. Their instructions were merely to 'rate' the distortion they heard on a scale of 0 to 5, using '0' to represent 'No distortion audible', and '5' to represent 'Distortion too severe to be tolerated in a B.B.C. programme'. In spite of the apparent vagueness of these instructions, statistical analysis reveals a remarkably small spread of opinion. This series of experiments was made by 31 observers selected from B.B.C. staff as having considerable experience and ability in judging programme quality.

The 'Average Crew Score', mentioned later, is the average of the rating figures recorded in each test presentation by all the observers who heard it—a number varying from 11 to 31 (see Table 1).

The Method of Distorting Programme Material

The method is given in detail in App. 1(c), and it is sufficient to make two points here; first, that although the loudness of the distorted signals followed the fluctuations of the original, the distortion percentage [in the sense explained in App. 1(c)] remained constant at the value chosen for the test item; and second, that the

TABLE 1
Arrangement of the Programme Tests

Group	*Number in Group*	*Programme Item Tested*	*Description of Programme Item*
A	6	1, 2, 3, 4	1. Piano solo
B	6	1, 2, 5, 6	2. Orchestra; The Secret Wedding; Cimarosa
C	5	1, 2, 5, 6	3. Military band
D	8	1, 2, 3, 4	4. Organ solo
E	6	1, 2, 3, 5	5. News and "Morning Story"
			6. Arabic news with Spanish story

Note: Since every group dealt with item 1, each of the 24 presentations of this item were voted on by 31 observers.
Similarly item 2 by 31 observers
item 3 by 20 observers.
item 4 by 14 observers.
item 5 by 17 observers.
item 6 by 11 observers.

distortion was generated by a single unbiased diode, the variation of distortion being produced by varying the dilution of the primary distortion. In the associated Fig. 9, the quantity S has the same significance as in the earlier section ('Arrangement of the Pulse Experiments'), and in App. 1(c), the conditions being virtually equivalent to half-wave rectification when $S = 10$.

The Grading-Function

Each test sequence contained 24 items selected from one of the programmes; to these each listener assigned a grading number. The degree of distortion applied to each item was chosen in irregular order, from a set of six or more values of S including zero; thus, there might be three or even four items for which S was set at the value 3, but these items would have only this distortion number in common, in every other respect they would be different—in frequency, loudness, and programme content.

Fig. 9 summarizes all the data. The choice of $(S)^2$ as the X co-ordinate was made after trying a number of alternatives. Because the plotted points fall around straight lines, the analysis of the next section is feasible.

The Sequential Diagram: the 'Drift' Effect

The next four diagrams, and the associated analysis, apply to the piano data summarized in Fig. 9(a); this has been chosen as the most representative because all 31 observers took part in the test.

Analysis develops on the following lines: The best-fitting line of Fig. 9(a) shows how the average-grading G_0 varies when distortion (as represented by S^2), varies. The diagram takes account of virtually nothing but distortion; the effects (if any) of all other factors are submerged and hidden by the averaging of the gradings assigned to different parts of the programme, and represented here by single plotted points.

Fig. 10(a) breaks down the averages of Fig. 9(a) into the component data G; these points, marked in serial order, show even greater dispersion about the G_0-line than do the averages in 9(a). The cause of this is sought, and when the dispersion has been examined in Figs. 10(b), 11 and 12, it is found to be due to three factors:

(a) A downward drift in the listeners' gradings as the test proceeds, due apparently to a clouding of their judgment under continual exposure to gross distortion [see Fig. 10(b)];

(b) an increase and decrease in the grading in sympathy with the changes of mean-frequency of the programme, shown in Fig. 11, which is modified by

(c) a regular re-establishment of a new frequency datum whenever the pitch-interval between items exceeds 2 tones (Fig. 12), or 3 tones (Fig. 13).

This analysis will now be taken in detail. The serial order of the 24 test conditions is shown in these figures by numbering them in pairs: 1, 1X, 2, 2X . . . up to 12, and 12X—so as to avoid large numbers. The X-axis of Figs. 10(b) and 11 is linear in time and makes allowance for the two 5-sec test items in each

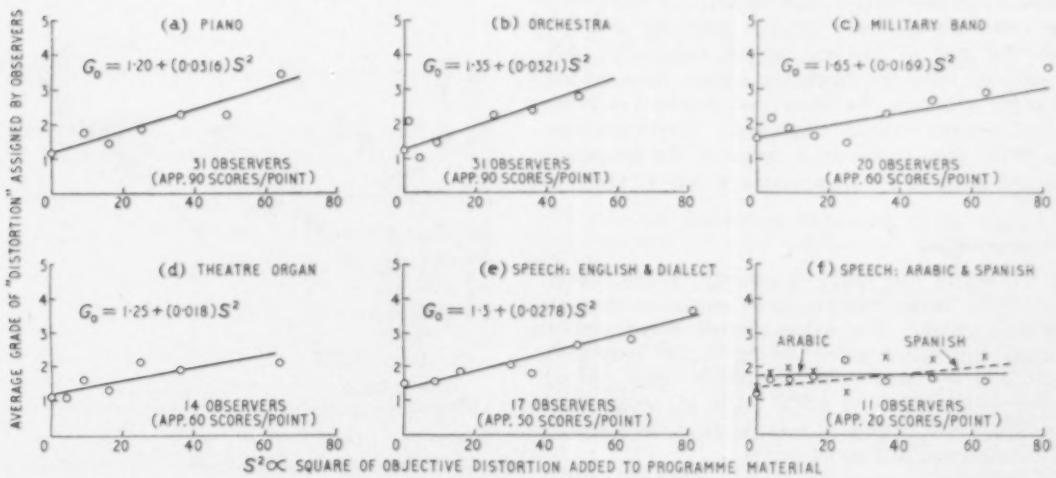


Fig. 9. *Distorted programme signals; typical grading functions. Assessment of the degree of distortion on an arbitrary subjective scale*

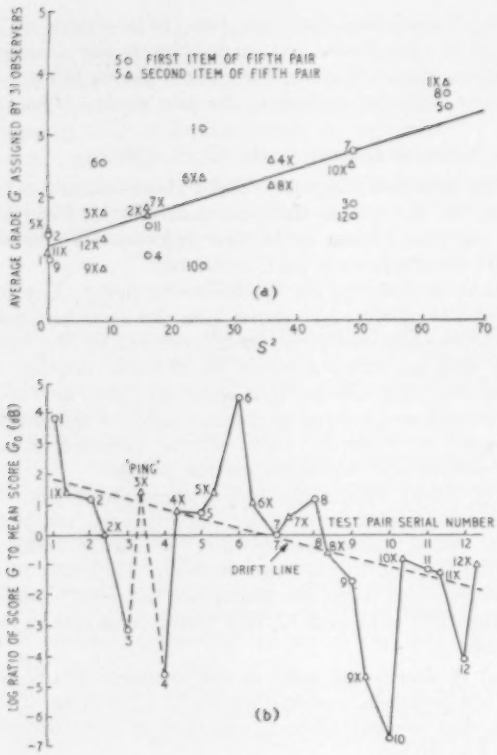


Fig. 10. *Evidence of drift in the audience reaction (piano test). (a) Deviations from the best fit in Fig. 9 (a); (b) sequential deviations*

pair, the 2-sec interval between items of a pair, and the 9-sec silent gap which follows.

To separate out the 'drift-effect', shown by the dotted line in Fig. 10(b), the departures of G from G_0, as deduced from Fig. 10(a), are plotted in time-sequence as the logarithm of the ratio (G/G_0). The drift-line is then interpolated by eye, under the restriction that it must pivot about the mid-point of the time axis. The downward slope indicates a 4-dB drop in the order-of-magnitude of the gradings assigned towards the end of the test-run as compared with those assigned near the beginning—thus, in a period of the order of 5-min, the observers' standard of judgment had become seriously degraded. The breakdown of Fig. 10(b) then proceeds in terms of the frequency factor.

The Frequency-Effect

Fig. 11 shows the mean frequency of test items. Unfortunately, measurements could not be made with any great precision. The values plotted were found by integrating the signal power falling in the passbands of high- and low-pass filters, emphasis being placed upon the crest values of each burst of programme signal. The mean frequency was taken at the point at which the filters passed equal power.

Since a change in mean frequency could be expected to produce a similar change of unpleasantness, Fig. 11 accounts to some extent for the low values of items 3 and 12 relative to the drift line, Fig. 10 (but some other explanation is needed for 8X). Clearly the artefact (marked 'Ping' at item 3X and produced by a valve in the recording amplifier) was responsible for the discrepancy here.

Now the 'drift' of Fig. 9(a), was disclosed by plotting the deviations of the scores G from the mean line G_0 and, in a similar way, the fine detail of the frequency effect can be revealed by plotting the deviations of the curve of Fig. 10(b) from the drift-line using, in this case, frequencies as the X-co-ordinate. Putting both deviations and frequency to a log scale, Fig. 12 results.

The frequency scale is marked off in semitones based upon 440 c/s as International-A. With the exception of three points, the data fall into two primary groups of points which lie, more or less, upon lines of 12-dB/octave slope. Not unexpectedly, all the points cannot be aligned, for the quantity plotted vertically has survived two separate subtraction processes, each of which will contribute their errors. The presentation sequence can be traced from point 1, near the top of Line 3, through 1X–2X–2 and 3, to the bottom of this line. In item 3X the artefact ('ping') was superimposed and this may account for the departure from the general pattern. Strictly, on the basis of its frequency, 3X belongs to line 2 of the pattern.

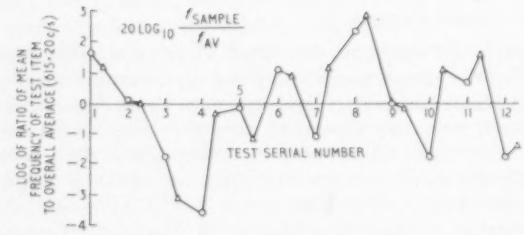


Fig. 11. *Measured mean frequency of 5-second items in piano programme*

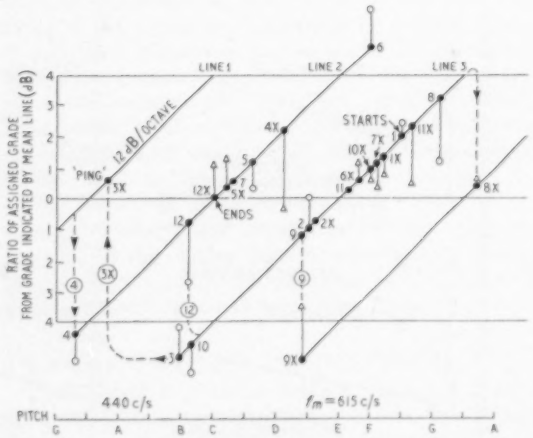


Fig. 12. *Frequency effect in piano programme. After due correction for drift of audience-reaction, unpleasantness is found to increase with pitch but in a cyclic way (31 observers)*

The basic pattern indicates a change in score if there is a change of frequency, provided that the range of frequency variation does not exceed ± 4 semitones; when that limit is exceeded the score/frequency relationship remains unchanged, but a new frequency-datum is established.

A single highly-skilled observer supplied the data for Fig. 13 which applies to the orchestral programme. This figure has the same general implications as Fig. 12. Notice that the score lines are 6 semitones apart while, in Fig. 12, the interval is 4 semitones.

Traces of a frequency effect were found in all the test data summarized in Fig. 9, except for (f), but the material was too meagre to justify full analysis.

Comments on Frequency and Contrast Effects

The frequency effect just dealt with shows a 12-dB/octave (i.e., square-law) relationship between score and frequency change (see Fig. 12); Fig. 9 shows a square-law relationship also between the average-crew score and the degree of objective distortion S; to complete the picture, the frequency effect illustrated in Fig. 5 was caused by an octave interval between the A and B signals under comparison, and the (admittedly uncertain) maximum value of the a.d.r. is not far short of 4·0, the square of the frequency ratio.

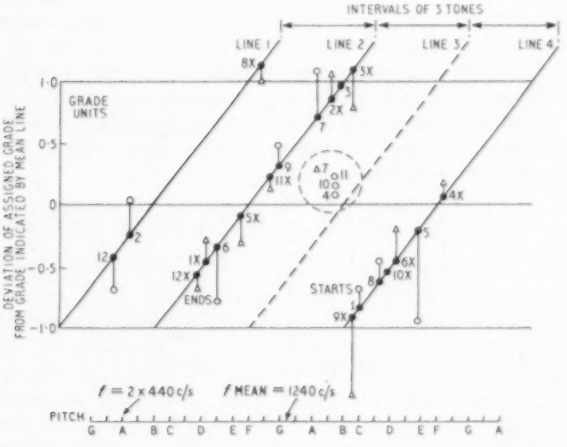


Fig. 13. *Frequency effect in the orchestral programme; this data was provided by a single highly-skilled observer*

The form of words adopted in the deviation-rate hypothesis is not specific on this point—the unpleasantness of distorted sounds is postulated as assessed '*in terms of* the time-rate-of-deviation of the signal from normality'; the data set out above supports that proposition in general terms, but suggests that the reference should be to the *square* of the deviation-rate.

Taken alone, the evidence of Fig. 12 is not strong; in Fig. 13 it will be noticed that the ordinate is expressed in linear instead of logarithmic units and in spite of this the data still falls into groups of lines. On the other hand, data calculated from (c), (d) and (e) in Fig. 9 fit the logarithmic presentation; they have not been illustrated because the pitch variations tend to fall into either one narrow group [in (e)] or several such groups.*

The audience's complete lack of discrimination in assessing the distortion gradings in (f) made frequency analysis impossible. The data provides evidence (although of a negative kind) for the 'contrast effect': it would appear that the distortion—although recognized as present (the average grading was near '2')—could not be specifically rated. There can be little doubt that this was because the languages were completely unknown to the audience; as 'normality' could not be recognized, neither could 'departure from normality', and this is the only acceptable measure of distortion. On this basis it was the lack of contrast between items which made specific rating of the distortion impossible.

REFERENCES

[1] G. von Gassler, *Frequenz*, 1955, Vol. 9, p. 15.
[2] H. J. von Braunmuhl and W. Weber, May 1937, *Akus. Zeit.*, Vol. 2, p. 135.

(*To be concluded*)

* The mathematically inclined reader will nevertheless have noticed that the logarithmic display of Fig. 12 would have to run into trouble had the G_0-line (the mean line of Fig. 9 (a)) fallen too near to zero. In addition he will have observed that the slope of the lines in Fig. 12 could have been altered merely by instructing the audience that unity (or any other numeral other than zero) was to be associated with the absence of distortion. Neither of these objections would have applied to the general form of the *linear* presentation, however.

ANGLO-U.S. AGREEMENT FOR SPECIAL PLASTICS

An agreement has been reached between Microcell and Emerson & Cuming Inc. of U.S.A. for the sale and distribution in the U.K. of the American Company's range of plastics, which have been specifically designed for the electronics industry.

The responsibility of covering Britain will be undertaken under the agreement by Microcell's Electronics Division, Blackwater, Camberley, Surrey.

Emerson & Cuming products include 'Eccosorb' which is a range of energy-absorbing materials covering the v.h.f., u.h.f. and microwave bands; 'Eccofoam'—a material supplied as a liquid, powder, or sheet and provides the user with plastic and ceramic foams which have known dielectric constants; 'Stycast'—a range of casting resins; and 'Ecco Reflector'—a material which is a passive reflector of radar energy and available in sizes up to 48 in. nominal diameter.

UNIVERSITY OF BIRMINGHAM GRADUATE COURSES

The Electrical Engineering Department of the University of Birmingham have announced two Graduate Courses for the year 1961–1962.

The Graduate Course in Information Engineering will be running for the fifth time and on the same general syllabus as in 1960–61. The course, which for 1961 was entitled "Theory of Electrical Machinery", will in future be known as "Electrical Machines: Electromechanical Energy Conversion and Automatic Control".

Both courses last twelve months from 1st October 1961 to 30th September 1962, and students who already hold a first degree of an approved university may qualify for the degree of M.Sc. by examination.

Enquiries regarding these courses should be addressed to the Graduate Course Supervisor, The University, Edgbaston, Birmingham, 15.

Audio-Frequency Comb-Filter

PRACTICAL SYSTEM USING A STRETCHED STRING

By L. H. Bedford, C.B.E.

The concept of a comb-filter (i.e., a filter having narrow pass or stop bands spaced at uniform frequency intervals), is one which occurs in various engineering contexts. The idea has not been extensively developed on a commercial basis because of the difficulties of practical realization. Recently, however, a problem arose which so strongly invited solution in terms of a comb filter that the writer found himself unable to resist the challenge. The result was the production of a decidedly elegant comb filter with extraordinarily little difficulty or apparatus. The experiment went so remarkably 'according to the book' that it is thought, if only for this reason, to be worth recording.

The problem which gave rise to this experiment is of no interest in itself, and indeed highly undignified, but needs to be stated as defining the problem. A tape recording had been made which owing to a technical fault was mutilated by mains hum. The possibility was therefore considered of performing a hum-stripping operation. The hum consisted of a nominal 50 c/s with a long series of odd harmonics.

The possibility of an ideal hum-stripping operation (i.e., one strictly without detriment to the music) depends on the fact that 50 c/s does not occur in the chromatic scale of standard pitch (i.e., based on A.440 c/s); the nearest note is G.49 c/s, so that we have a clear 2% frequency separation between wanted and unwanted signals. If we demand 20-dB attenuation for the hum components and not more than 3 dB at 2% frequency displacement, this implies, for a single resonant circuit, a Q of 275 approx.

Such Q values at low frequencies are scarcely obtainable with LC networks but are readily available in mechanical systems; e.g., a stretched string. Further, the string has modes substantially at the harmonic frequencies and so presents the possibility of providing the desired series of attenuation bands in one fell swoop.

The arrangement which translates this idea into practice is shown in Fig. 1. A (non-magnetic) metallic string, length l, is stretched between bridges so as to resonate in its fundamental mode at 50 c/s. Across the centre section of the string, length βl, is straddled a magnetic field, strength B. If alternating current is passed through the string its motion introduces a 'dynamic impedance' Z_d over and above its d.c. resistance Z_0. Z_d becomes appreciable only in the vicinity of the resonant frequencies at which quite high values of Z_d/Z_0 are obtainable. Fig. 2 is a numerical example showing the behaviour of the string dynamic impedance, and the resultant Z_f.

The above suggests the circuit of Fig. 3 as a promising arrangement. The amplifier is conveniently of the Williamson type allowing a large amount (say, 26 dB) of negative feedback. The normal feedback circuit is substituted by a potentiometer network consisting of a fixed resistance R_n and the string impedance $Z_0 + Z_d$. Off resonance this amounts merely to a resistive-potentiometer network which is adjusted to give the normal amount of negative feedback. For a frequency corresponding to a string resonance the negative feedback is increased, thus producing a corresponding attenuation in the system.

In order to secure 20 dB of extra negative feedback at resonance it is necessary for Z_d to be some nine times Z_0 and for R_n to be high compared to Z_d.

Note that we are now contemplating a total negative feedback of 46 dB at the string resonances, which is at first rather frightening. The situation is saved, however, by the restricted frequency range over which the resonances are effective.

To obtain a value of $\frac{Z_d}{Z_0}$ of 9 is rather more than can conveniently be managed. The equivalent result is therefore obtained by an artifice. Since Z_d cannot easily be increased it is necessary to find a means of reducing the effective value of Z_0. This is done by the use of the positive-feedback network shown in Fig. 3. If, for example, we apply 6 dB of positive feedback the *effective* value of Z_0 is halved. In general, we can modify the effective value of Z_0 to Z_0' where $0 > Z_0' < Z_0$. [Ref. Fig. 2(b)].

The design of the string system is determined by the following equations:

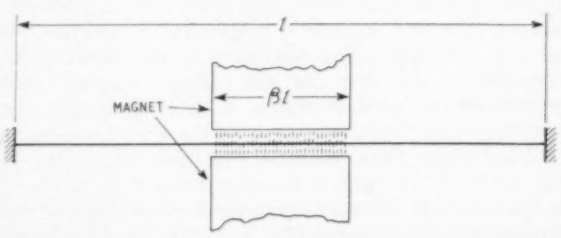


Fig. 1. *The non-magnetic stretched-string and magnet arrangement*

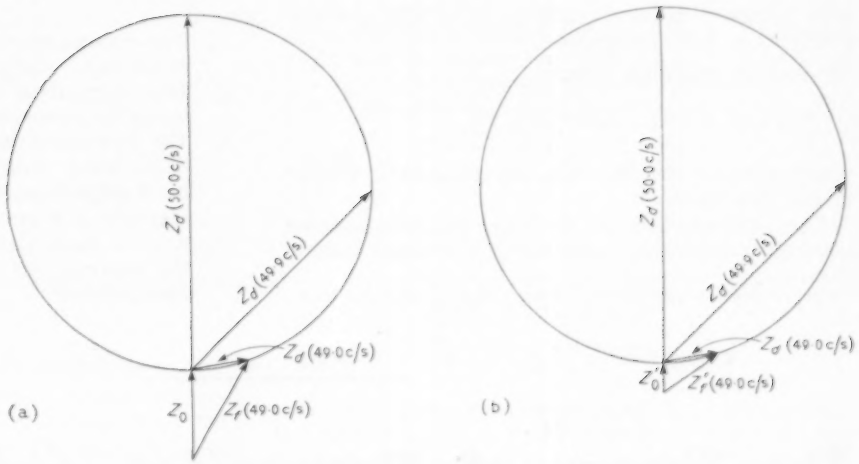


Fig. 2. *(a) A numerical example showing the behaviour of the string dynamic impedance and the resultant Z_f; (b) shows that by applying positive feedback Z_0 is reduced to Z_0' and, therefore, Z_d/Z_0 is increased*

$$Z_d = \frac{4}{\pi^2} QB^2 \frac{1}{\omega} \frac{l}{m} \left(\frac{1}{n} \sin n\beta \frac{\pi}{2} \right)^2 \quad .. \quad .. \quad (1)$$

whence

$$\frac{Z_d}{Z_0} = \frac{4}{\pi^2} QB^2 \frac{1}{n\omega_1} \frac{1}{\rho\sigma} \left(\frac{1}{n} \sin n\beta \frac{\pi}{2} \right)^2 \quad .. \quad .. \quad (2)$$

and for $n\beta \ll 1$,

$$\frac{Z_d}{Z_0} \approx QB^2 \frac{1}{n\omega_1} \frac{1}{\rho\sigma} \beta^2 \quad .. \quad .. \quad .. \quad (2a)$$

Here

- Z_0 is the d.c. resistance of the string
- Z_d is the 'dynamic impedance' due to motion at resonance
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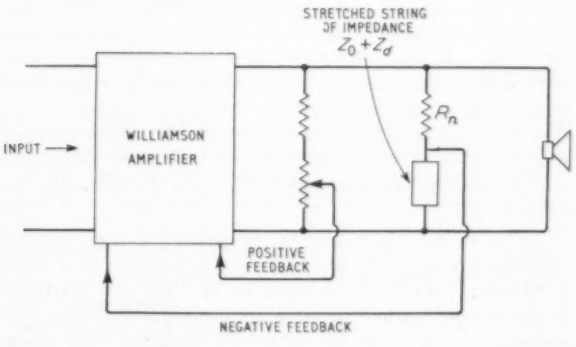

Fig. 3. *The arrangement used to provide attenuation at the frequency corresponding to the string resonance*

For the 'run out' of the mode frequencies from the strict harmonic series the relevant formula is

$$\frac{f_n}{f_1} = n \left\{ 1 + \frac{1}{32} \pi^2 \left(\frac{d}{l} \right)^2 \varkappa^{-1} (n^2 - 1) \right\} \quad .. \quad (3)$$

This is Rayleigh's second-order stiffness correction for a round wire of diameter d tensioned to strain $\varkappa$. There

TABLE 1

Mode No. n	Frequency of max. atten. (c/s)	Max. attenuation (dB)	Bandwidth, cents, between points of 6-dB attenuation	Inferred Q of string in nth mode
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7	7 × 49 + 6·5 ,,	18·7	13	710
9	9 × 49 + 8·0 ,,	18·7	8	1140
11	11 × 49 + 7·0 ,,	17·5	6	1400
13	13 × 49 + 6·8 ,,	16·5	4·5	1700
15	15 × 49 + 6·8 ,,	16·5	4·5	1700
17	17 × 49 + 6·5 ,,	15·6	3·0	2200
19	19 × 49 + 7·0 ,,	14·5	2·0	1600

† N.B. For these measurements the string was let down to 49 c/s, approx., to suit the calibration of the measuring instrument.

may also be other minor effects due to air loading and finite rigidity of termination but these are not calculable.

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In any case, it must be admitted that Equ. (2) conceals certain vital information by merely quoting a Q value without further analysis. In point of fact so many factors enter into this that only the experimental approach is expedient. This has been adopted

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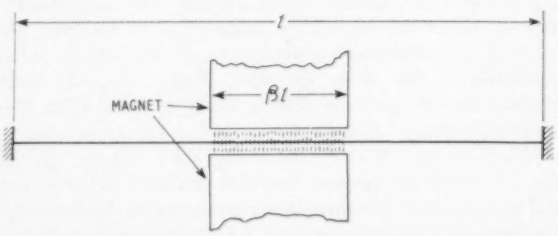

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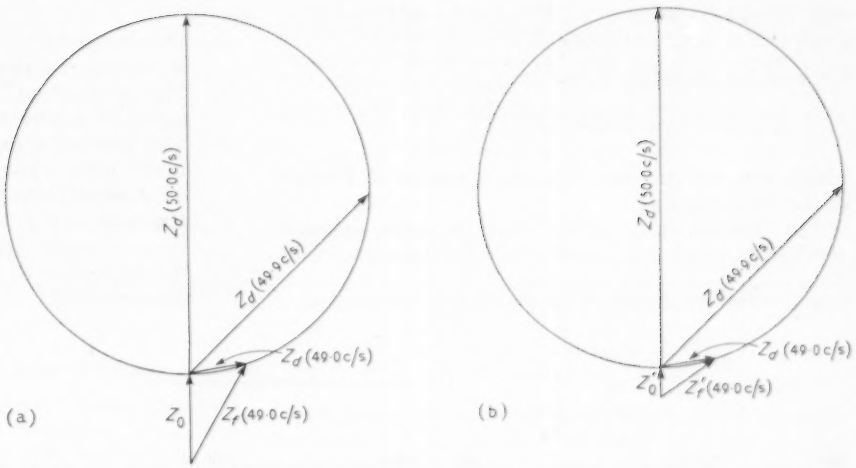

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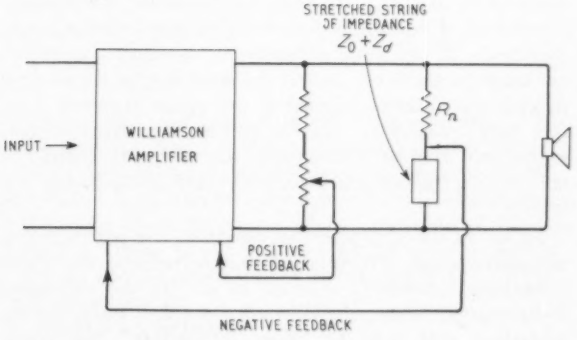

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with arbitrary values $l = 152$ cm, $\beta = 0{\cdot}06$ and $d = 0{\cdot}008$ in. $= 0{\cdot}02$ cm approx.

We find for phosphor bronze

$$\frac{f_n}{f_1} = n\{1 + 3{\cdot}24 \times 10^{-6}(n^2 - 1)\},$$

which is to say that the 'run out' amounts to 1 cent* at the 13th harmonic.

The characteristics of the experimentally-achieved comb filter are shown in the table. The salient features are:—

1. The extreme sharpness of the 'teeth', especially in the higher modes.
2. The attenuation is maintained in the higher modes by a providential increase in the Q values.
3. The 'run out' of the mode frequencies is extremely small, being rather less that that predicted by the Raleigh formula, and not of the same form. However, it is probable that the tabulated deviations of mode frequencies reflect no more than the experimental inaccuracies of the (difficult) measurement.

* 1 cent = 1 hundredth of a semitone = 0·0595% frequency increment.

The Theory of Waveguides and Cavities

2: EXAMPLES OF WAVEGUIDES AND DISCUSSION OF SPECIAL POINTS

By R. A. Waldron, M.A. (Cantab.), A. Inst. P., A.M. Brit I.R.E.*

SUMMARY. *The general approach described in the first article of this series is applied to two examples, the guide consisting of a metal tube of circular cross-section enclosing air, and the coaxial line, taking the core material to be a general dielectric instead of a perfect conductor. This approach enables losses to be treated, and permits the investigation of the E_{p0} modes, which are normally regarded as non-existent but which can in fact propagate with difficulty in the waveguide. The transmission-line modes of the coaxial line are related to the E_{p0} modes of the guide. Finally, the nature of E and H modes, power flow, and the impedance concept are discussed in relation to general waveguides.*

In the first article of this series[1], a general approach was described for examining waveguides and cavities theoretically. In the present article, the method will be used to examine two examples. One is the guide consisting of a tube of metal enclosing air, the other the coaxial line. We shall take more general models of these than are usually taken; the treatment is, of course, more complicated, but it leads to more general results than can usually be obtained, from which a deeper insight into the structure of the mode spectrum can be obtained.

The method usually given in the text books for these problems is to take the metal to be a perfect conductor, and the air to have the same properties as vacuum. The method we shall adopt gives a result which reduces to the usual one when air is replaced by vacuum and metal by a perfect conductor, but which in its general form enables us to take account of the effect of the finite conductivity of the metal and the loss in the air. It will also enable us to bring out some points concerning the modes which are not apparent from the restricted treatment.

Outline of the Method

The method was fully described in the previous article, and details should be sought there. However, to enable the present article to be read without referring back, a brief outline will be given here.

Maxwell's equations are taken as starting point. From these a pair of wave equations are derived, containing E_z and H_z as independent variables, z being the direction of the guide axis. Solution of the wave equations gives E_z and H_z in terms of certain arbitrary constants and the phase constant. The other field components are obtained from Maxwell's equations in terms of E_z and H_z. By applying boundary conditions, a number of equations for the arbitrary constants are obtained. If the same number of boundary conditions are taken as there are constants, these can be eliminated to give an equation involving the phase constant β as the only unknown; this is called the characteristic equation. If β is determined, the arbitrary constants can be determined and hence the field components are known.

As far as the characteristic equation the procedure is straightforward. The analysis may be tedious, but there is nothing inherently difficult in it. In the examples to be treated below, the derivations of the characteristic equations will only be given in outline; the main

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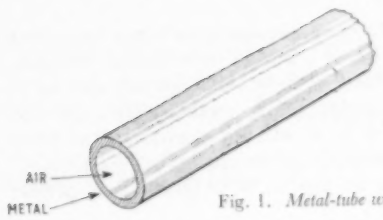

Fig. 1. *Metal-tube waveguide*

interest lies in discussing the characteristic equations themselves.

The Waveguide of Circular Cross-Section

The waveguide to be treated here consists of a metal tube of circular cross-section, containing air. Two models will be considered here. The first is the usual one treated in the text books; the guide is considered as a homogeneous simple perfect (HSP) guide. By a simple guide is meant one in which only scalar media are present, such as metals, dielectrics, vacuum, or air, but not ferrites or plasmas. A perfect guide is one in which all the media are lossless. A guide is described as homogeneous if the energy of the waves is confined to the interior of a boundary surface that encloses a single homogeneous medium; for this to be the case, the medium outside the surface must be a perfect conductor. For this model the metal is replaced by a perfect conductor and the space inside it is regarded as a vacuum. The second model we shall take is more general. Firstly, we shall take the air to be a lossy scalar dielectric, which in fact it is, and secondly we shall consider the metal to be a metal. The approximation we shall make is to take the metal to extend to infinity. Since the thickness of the metal in any practical case is very large compared with the skin depth, the difference between this model and the actual physical waveguide is minute.

The Homogeneous Simple Perfect Model

For this model we take the permittivity and permeability of the medium to be those of free space, i.e., ϵ_0 and μ_0. It is readily shown from Maxwell's equations that the wave equations are

$$\left.\begin{aligned}(\nabla_t^2 + k_0^2 H_z) &= 0\\ (\nabla_t^2 + k_0^2 E_z) &= 0\end{aligned}\right\} \quad .. \quad .. \quad .. \quad .. \quad (1)$$

where $k_0^2 = \omega^2\epsilon_0\mu_0 - \beta^2 = \omega^2\epsilon_0\mu_0\,(1 - \nu^2)$

$$= \frac{4\pi^2}{\lambda_0^2}\,(1 - \nu^2) \quad .. \quad .. \quad .. \quad (2)$$

and $\nabla_t^2 = \nabla^2 - \partial^2/\partial z^2$, i.e., ∇_t^2 is what remains if, on writing out ∇^2, the Laplacian operator, in full in the appropriate co-ordinate system (in this case the cylindrical system r, θ, z), and then dropping $\partial^2/\partial z^2$, Equ. (2) can be regarded as a definition of the normalized phase constant ν; more explicitly, $\nu = \lambda_0\beta/2\pi = \lambda_0/\lambda_g$, where λ_0 and λ_g are the wavelengths in free space and in the guide, respectively.

The solutions of Equs (1) are

$$\left.\begin{aligned}E_z &= A\mathrm{J}_p\,(k_0 r)\,e^{-jp\theta}\,e^{-j\beta z}\,e^{j\omega t}\\ H_z &= B\mathrm{J}_p\,(k_0 r)\,e^{-jp\theta}\,e^{-j\beta z}\,e^{j\omega t}\end{aligned}\right\} \quad .. \quad .. \quad (3)$$

where p is an integer and $\mathrm{J}_p(x)$ is the Bessel function of the first kind of order p and argument x. The Bessel function of the second kind does not occur because at $r = 0$ this goes to infinity, while the fields must remain finite. The other field components are given by

$$\left.\begin{aligned}k_0^2 H_r &= \frac{2\pi}{\lambda_0}\left\{\frac{p}{r}\sqrt{\frac{\epsilon_0}{\mu_0}}\,E_z - j\nu\,\frac{\partial H_z}{\partial r}\right\}\\ k_0^2 H_\theta &= -\frac{2\pi}{\lambda_0}\left\{j\sqrt{\frac{\epsilon_0}{\mu_0}}\,\frac{\partial E_z}{\partial r} + \frac{p\nu}{r}\,H_z\right\}\\ k_0^2 E_r &= -\frac{2\pi}{\lambda_0}\left\{j\nu\,\frac{\partial E_z}{\partial r} + \frac{p}{r}\sqrt{\frac{\mu_0}{\epsilon_0}}\,H_z\right\}\\ k_0^2 E_\theta &= \frac{2\pi}{\lambda_0}\left\{-\frac{p\nu}{r}\,E_z + j\sqrt{\frac{\mu_0}{\epsilon_0}}\,\frac{\partial H_z}{\partial r}\right\}\end{aligned}\right\} \quad (4)$$

It is frequently stated in text books that it follows from the form of Equs (1) that two series of modes can be considered, the E modes having $E_z \neq 0$ and $H_z = 0$, and the H modes, having $E_z = 0$ and $H_z \neq 0$. The point was first made by Lord Rayleigh, and has been repeated by nearly every subsequent writer on waveguide theory. It will be seen from the following treatment that there are two series of modes; however, this is a fortunate accident, which does not follow from the fact that Equs (1) contain either E_z or H_z, but not both in the same equation. We shall return to this point later. (See section "E Modes and H Modes".)

Let us now apply boundary conditions. There are two arbitrary constants, so we use two conditions. Firstly, at $r = a$, a being the radius of the guide surface, E_z must vanish, since it is parallel to a perfect conductor. Thus from the first of Equs (3),

$$\mathrm{J}_p(k_0 a) = 0 \quad .. \quad .. \quad .. \quad .. \quad .. \quad (5)$$

Secondly, at $r = a$, E_θ must vanish. From Equs (4), this requires

$$-\frac{p\nu}{r}\,E_z + j\sqrt{\frac{\mu_0}{\epsilon_0}}\,\frac{\partial H_z}{\partial r} = 0$$

Since E_z vanishes at $r = a$, this reduces to

$$\partial H_z/\partial r = 0$$

and so from the second of Equ. (3),

$$\mathrm{J}_p'(k_0 a) = 0 \quad .. \quad .. \quad .. \quad .. \quad .. \quad (6)$$

Thus there are two series of modes, characterized by Equs. (5) and (6). These equations are to be solved for k_0, and then ν is obtained from Equs (2). The solution

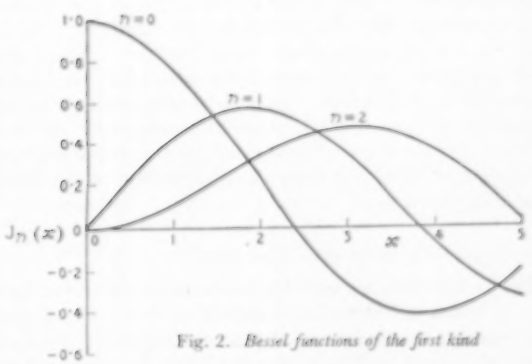

Fig. 2. *Bessel functions of the first kind*

presents no difficulty, since Bessel functions are extensively tabulated.

It will be noticed that the first boundary condition can be satisfied regardless of the value of B, and therefore, in particular, when $B = 0$; and the second boundary condition can be satisfied regardless of the value of A, and therefore, in particular, when $A = 0$. Thus Equ. (5) refers to waves which can exist with $H_z = 0$, and these we call E waves; while Equ. (6) refers to H waves, having $E_z = 0$.

It is merely by accident that waves having E_z or H_z zero exist in the HSP guide. No great significance attaches to the fact; it renders the theoretical treatment of these guides slightly simpler than would otherwise be the case, but it should not be allowed to colour our ideas about waveguides in general, for in a guide which is not HSP, while there are still two series of modes, they are not characterized by the vanishing of E_z or H_z. This point will be discussed more fully in a later section (E Modes and H Modes).

The mode spectrum is readily obtainable from Equs (5) and (6). Having assigned a value to p, the qth solution of Equs. (5) or (6) enables the phase constant of

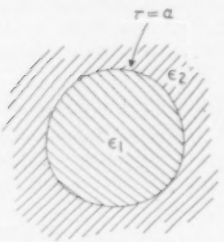

Fig. 3. *Simple waveguide with two media. When* $\epsilon_2' = 0$, $\epsilon_2'' \to \infty$, $\epsilon_1' = 1$, $\epsilon_1'' \ll 1$, *this approximates to the metal-tube waveguide of Fig.* 1

the E_{pq} or H_{pq} mode, as the case may be, to be found. The zeros of these equations all occur for real positive values of $k_0 a$, i.e., of $\sqrt{1 - \nu^2}$. Thus ν is either real, lying between 0 and 1, or imaginary. If real, the mode propagates in the guide; if imaginary, the guide is cut off for that mode. $\nu = 0$ is the cut-off condition. Then Equs (5) and (6) reduce respectively to

$$J_p(2\pi a/\lambda_0) = 0 \quad .. \quad .. \quad .. \quad .. \quad (7)$$

$$\text{and } J_p'(2\pi a/\lambda_0) = 0 \quad .. \quad .. \quad .. \quad .. \quad (8)$$

which can be solved for λ_0, the cut-off value of the free-space wavelength.

From the above treatment, the mode spectrum of the HSP guide can be obtained, and this is a very close approximation to that of the actual physical waveguide. The treatment has the advantage of simplicity; however, it has compensating disadvantages. Since the HSP guide is lossless, there is no possibility of taking losses into consideration; having investigated the mode spectrum of the actual physical waveguide by this method, losses can only be treated by the application of perturbation theory, a topic to be discussed in the third article of this series.

The Simple Waveguide with Two Media

For this model, we take the medium inside the surface $r = a$ to have a dielectric constant $\epsilon_1 = \epsilon_1' - j\epsilon_1''$ and the medium outside to have a dielectric constant $\epsilon_2 = \epsilon_2' - j\epsilon_2''$. The permeability is everywhere μ_0. We shall not impose any restrictions as yet on ϵ_1 and ϵ_2, except that ϵ_1', ϵ_2', ϵ_1'', and ϵ_2'' are all real, but subsequently we shall take ϵ_1' to be unity for air, and ϵ_2' to be zero, ϵ_2'' very large, for a metal. Also, for the medium outside $r = a$, it will be convenient to write $\mu_0 \mu_2$ for the permeability, although in fact μ_2 is unity; this will enable us to put $\mu_2 = 0$ later for a perfect conductor.

We proceed by analogy with the earlier section (The Homogeneous Simple Perfect Model). The wave equations are:

$$\text{for } r < a \quad (\nabla_t^2 + k_1^2)\begin{Bmatrix} E_z \\ H_z \end{Bmatrix} = 0 \quad .. \quad .. \quad (9)$$

$$\text{for } r > a \quad (\nabla_t^2 + k_2^2)\begin{Bmatrix} E_z \\ H_z \end{Bmatrix} = 0 \quad .. \quad .. \quad (10)$$

$$\left.\begin{aligned} \text{where } k_1^2 &= \omega^2\epsilon_0\mu_0(\epsilon_1 - \nu^2) \\ k_2^2 &= \omega^2\epsilon_0\mu_0(\epsilon_2\mu_2 - \nu^2) \end{aligned}\right\} \quad .. \quad .. \quad .. \quad (11)$$

As before, the solutions for medium 1 ($r<a$) are

$$\left.\begin{aligned} E_z &= AJ_p(k_1 r)\, e^{-jp\theta}\, e^{-j\beta z}\, e^{j\omega t} \\ H_z &= BJ_p(k_1 r)\, e^{-jp\theta}\, e^{-j\beta z}\, e^{j\omega t} \end{aligned}\right\} \quad .. \quad .. \quad .. \quad (12)$$

and for medium 2 ($r>a$)

$$\left.\begin{aligned} E_z &= CH_p(k_2 r)\, e^{-jp\theta}\, e^{-j\beta z}\, e^{j\omega t} \\ H_z &= DH_p(k_2 r)\, e^{-jp\theta}\, e^{-j\beta z}\, e^{j\omega t} \end{aligned}\right\} \quad .. \quad .. \quad .. \quad (13)$$

Here $H_p(x)$ is used to denote the Hankel function of the second kind, given by

$$H_p(x) = J_p(x) - jY_p(x) \quad .. \quad .. \quad .. \quad (14)$$

for integral values of p. This form of solution is chosen to satisfy the boundary conditions at infinity, which are that the fields shall approach zero as r approaches infinity, and that at large finite distances energy travels outwards from the system, but not inwards. The other field components in medium 2 are given by

$$\left.\begin{aligned} k_2^2 H_r &= \frac{2\pi}{\lambda_0}\left\{\frac{\epsilon_2 p}{r}\sqrt{\frac{\epsilon_0}{\mu_0}}E - j\nu\frac{\partial H_z}{\partial r}\right\} \\ k_2^2 H_\theta &= -\frac{2\pi}{\lambda_0}\left\{j\sqrt{\frac{\epsilon_0}{\mu_0}}\epsilon_2\frac{\partial E_z}{\partial r} + \frac{p\nu}{r}H_z\right\} \\ k_2^2 E_r &= -\frac{2\pi}{\lambda_0}\left\{j\nu\frac{\partial E_z}{\partial r} + \mu_2\frac{p}{r}\sqrt{\frac{\mu_0}{\epsilon_0}}H_z\right\} \\ k_2^2 E_\theta &= \frac{2\pi}{\lambda_0}\left\{-\frac{p\nu}{r}E_z + j\mu_2\sqrt{\frac{\mu_0}{\epsilon_0}}\frac{\partial H_z}{\partial r}\right\} \end{aligned}\right\} \quad (15)$$

For medium 1, the field components are obtained from Equs (15) by replacing k_2 and ϵ_2 by k_1 and ϵ_1, and μ_2 by unity.

There are four arbitrary constants, and so four boundary conditions are needed. These are that at $r = a$, E_z, H_z, E_θ, and H_θ are continuous. We thus obtain

$$AJ_p(k_1 a) - CH_p(k_2 a) = 0 \quad .. \quad .. \quad .. \quad (16)$$

$$BJ_p(k_1 a) - DH_p(k_2 a) = 0 \quad .. \quad .. \quad .. \quad (17)$$

$$\left.\begin{aligned} &PAJ_p(k_1 a) - jR_1R_2^2\sqrt{\frac{\mu_0}{\epsilon_0}}BJ_p'(k_1 a) \\ &\quad + jR_2R_2^2\sqrt{\frac{\mu_0}{\epsilon_0}}\mu_2 DH_p'(k_2 a) = 0 \end{aligned}\right\} \quad (18)$$

$$jR_1 R_1^2 \sqrt{\frac{\epsilon_0}{\mu_0}}\,\epsilon_1 A \mathrm{J}_p'(k_1a) + PB\mathrm{J}_p(k_1a) - jR_2 R_1^2 \sqrt{\frac{\epsilon_0}{\mu_0}}\,\epsilon_2 C \mathrm{H}_p'(k_2a) = 0 \quad .. \quad (19)$$

$$\left.\begin{aligned} \text{where} \quad P &= \frac{pv}{2\pi a/\lambda_0}(\epsilon_2\mu_2 - \epsilon_1) \\ R_1 &= k_1/\omega\sqrt{\epsilon_0\mu_0} \\ R_2 &= k_2/\omega\sqrt{\epsilon_0\mu_0} \end{aligned}\right\} \quad .. \quad .. \quad (20)$$

These four equations in A, B, C, and D are consistent if the determinant of the coefficients vanishes, i.e.

$$\begin{vmatrix} \mathrm{J}_p(k_1a) & 0 & \mathrm{H}_p(k_2a) & 0 \\ 0 & \mathrm{J}_p(k_1a) & 0 & \mathrm{H}_p(k_2a) \\ P\mathrm{J}_p(k_1a) & jR_1R_2^2\sqrt{\frac{\mu_0}{\epsilon_0}}\mathrm{J}_p'(k_1a) & 0 & jR_2R_1^2\sqrt{\frac{\mu_0}{\epsilon_0}}\mu_2\mathrm{H}_p'(k_2a) \\ jR_1R_2^2\sqrt{\frac{\epsilon_0}{\mu_0}}\epsilon_1\mathrm{J}_p'(k_1a) & -P\mathrm{J}_p(k_1a) & jR_2R_1^2\sqrt{\frac{\epsilon_0}{\mu_0}}\epsilon_2\mathrm{H}_p'(k_2a) & 0 \end{vmatrix} = 0$$

This is the characteristic equation. Multiplying out, we obtain

$$\left\{\begin{matrix} \epsilon_2(\epsilon_1-\nu^2)\sqrt{\mu_2\epsilon_2-\nu^2}\,\mathrm{J}_p(k_1a)\,\mathrm{H}_p'(k_2a) \\ -\epsilon_1(\epsilon_2\mu_2-\nu^2)\sqrt{\epsilon_1-\nu^2}\,\mathrm{J}_p'(k_1a)\,\mathrm{H}_p(k_2a) \end{matrix}\right\}\left\{\begin{matrix} \mu_2(\epsilon_1-\nu^2)\sqrt{\mu_2\epsilon_2-\nu^2}\,\mathrm{J}_p(k_1a)\,\mathrm{H}_p'(k_2a) \\ -(\epsilon_2\mu_2-\nu^2)\sqrt{\epsilon_1-\nu^2}\,\mathrm{J}_p'(k_1a)\,\mathrm{H}_p(k_2a) \end{matrix}\right\} - \frac{p^2\nu^2}{4\pi^2 a^2/\lambda_0^2}(\epsilon_2\mu_2-\epsilon_1)^2\mathrm{J}_p^2(k_1a)\,\mathrm{H}_p^2(k_2a) = 0 \quad .. \quad .. \quad (21)$$

This equation applies to any two scalar dielectric media separated by the boundary surface $r = a$; it is only a question of substituting the appropriate values for ϵ_1 and ϵ_2. For the problem we are considering, we now write

$$\epsilon_1 = \epsilon_1' - j\epsilon_1'' = 1 - j\epsilon_1''$$

for air, and

$$\epsilon_2 = -j\epsilon_2''$$

for a metal, with $\epsilon_1'' \ll 1$ and ϵ_2'' very large, but not infinite. This will enable us to take into account the losses due to the dielectric loss of the air and the finite conductivity of the metal. In principle, Equ. (21) could be solved exactly, but it is sufficiently accurate for practical purposes, and far easier, to obtain a first-order Taylor expansion for ϵ_1'', and this is what we shall do here. The solution gives β as a complex quantity, $\beta' - j\beta''$. Now, the propagation constant is

$$\gamma = \alpha - j\beta$$

If we write

$$\gamma = j(\beta' - j\beta''),$$

it is seen that $\beta'' = \alpha$, i.e., the imaginary part of the phase constant is the attenuation constant.

It is convenient to take the dielectric and wall losses separately. First we treat the dielectric loss, and neglect all but the highest-order terms in ϵ_2. The product of the two brackets in Equ. (21) is of order $\epsilon_2^2\sqrt{\epsilon_2}$, while the final term (in $p^2\nu^2$) is of order ϵ_2^2 and so may be neglected. Thus either of the two brackets may be separately equated to zero. In the first bracket, the first term is of order $\epsilon_2\sqrt{\epsilon_2}$, the second of order ϵ_2. For large ϵ_2, we take only the first term, giving

$$\mathrm{J}_p(k_1a)\,\mathrm{H}_p'(k_2a) = 0 \quad .. \quad .. \quad .. \quad .. \quad (22)$$

In the second bracket, only the second term survives, giving

$$\mathrm{J}_p'(k_1a)\,\mathrm{H}_p(k_2a) = 0 \quad .. \quad .. \quad .. \quad .. \quad (23)$$

For large ϵ_2, $k_2a \doteqdot \frac{2\pi a}{\lambda_0}\cdot\sqrt{-j\epsilon_2''}$, lying near a line at 45° to the real axis in the complex k_2a plane. For such values of argument, Hankel functions and their derivatives have no zeros. Hence from Equs. (22) and (23) we have

$$\left.\begin{aligned} \mathrm{J}_p(k_1a) &= 0 \\ \mathrm{J}_p'(k_1a) &= 0 \end{aligned}\right\} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (24)$$

respectively. Now,

$$k_1a = \frac{2\pi a}{\lambda_0}\sqrt{1-\nu^2-j\epsilon_1''} = \frac{2\pi a}{\lambda_0}\sqrt{1-\nu^2}\left\{1 - \frac{j\epsilon_1''/2}{1-\nu^2}\right\} = k_0a\left\{1 - \frac{j\epsilon_1''/2}{1-\nu^2}\right\}$$

Making this substitution in Equs (24) and expanding by Taylor's theorem, we obtain respectively

$$\left.\begin{aligned} \mathrm{J}_p(k_0a) - \frac{j\epsilon_1''\pi a/\lambda_0}{\sqrt{1-\nu^2}}\,\mathrm{J}_p'(k_0a) &= 0 \\ \mathrm{J}_p'(k_0a) - \frac{j\epsilon_1''\pi a/\lambda_0}{\sqrt{1-\nu^2}}\,\mathrm{J}_p''(k_0a) &= 0 \end{aligned}\right\} \quad .. \quad (25)$$

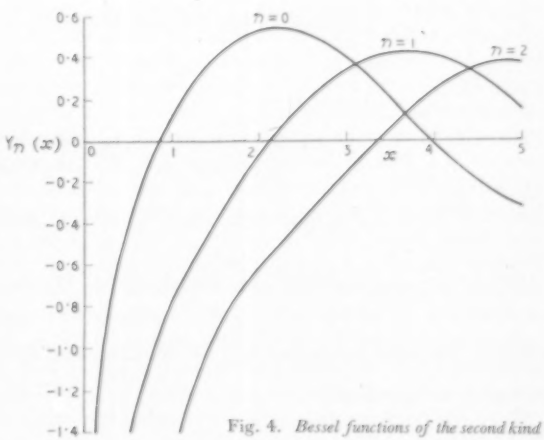

Fig. 4. *Bessel functions of the second kind*

On putting $\epsilon_1'' = 0$, these reduce to Equs (5) and (6), permitting a comparison of the mode spectrum of the more general model with that of the HSP guide. If a solution of, say, Equ. (5) is known, this may be taken as a first approximation to the solution of the first of Equs (25). Let the solution of Equ. (5) be ν_0, and let the corresponding solution of the first of Equs (25) be $\nu_0 + \delta\nu$. We can write

$$\sqrt{1 - (\nu_0 + \delta\nu)^2} \doteqdot \sqrt{1 - \nu_0^2}\left\{1 - \frac{\nu_0\,\delta\nu}{1 - \nu_0^2}\right\}$$

Then the first of Equs (25) becomes

$$J_p(k_0a) - \frac{2\pi a}{\lambda_0}\,\frac{\nu_0\delta\nu}{\sqrt{1-\nu_0^2}}\,J_p'(k_0a) - \frac{\pi a}{\lambda_0}\,\frac{j\epsilon_1''}{\sqrt{1-\nu_0^2}}\,J_p'(k_0a) = 0 \quad .. \quad (26)$$

By virtue of Equ. (5), the first term vanishes. Hence

$$\delta\nu = -j\epsilon_1''/2\nu_0 \quad .. \quad .. \quad .. \quad .. \quad .. \quad (27)$$

and this is the attenuation constant for the dielectric loss for E modes. The same result is obtained in the same way for H modes from the second of Equs (25).

For the wall loss, we take ϵ_1'' to be zero, ϵ_1' to be unity, and consider the highest and second highest terms in ϵ_2. Thus we obtain from Equ. (21)

$$J_p(k_0a)\,J_p p'(k_0a)\,H_p(k_2a)\,H_p'(k_2a) - e^{j\pi/4}\,\frac{\sqrt{\mu_2/\epsilon_2''}}{\sqrt{1-\nu^2}}\,J_p'^2(k_0a)\,H_p^2(k_2a) - \epsilon_1 e^{j\pi/4}\sqrt{\frac{\mu_2}{\epsilon_2}}\sqrt{1-\nu^2}\,J_p^2(k_0a)\,H_p'^2(k_2a) - e^{j\pi/4}\,\frac{\sqrt{\mu_2/\epsilon_2''}}{(1-\nu^2)^{3/2}}\,J_p^2(k_0a)\,H_p^2(k_2a) = 0 \quad .. \quad (28)$$

In the limit as $\epsilon_2'' \to \infty$, i.e., for a perfect conductor, only the first term of Equ. (28) survives, and gives Equs (5) and (6). For the loss, write $\nu = \nu_0 + \delta\nu$ as before, and consider E modes. The treatment for H modes is analogous. For Equ. (28) we may write, dividing by $J_p'(k_0a)\,H_p(k_2a)\,H_p'(k_2a)$,

$$J_p(k_0a) - e^{j\pi/4}\,\frac{\sqrt{\mu_2/\epsilon_2''}}{\sqrt{1-\nu^2}}\,J_p'(k_0a)\,H_p(k_2a)/H_p'(k_2a) - \epsilon_1 e^{j\pi/4}\sqrt{\frac{\mu_2}{\epsilon_2''}}\sqrt{1-\nu^2}\,J_p^2(k_0a)\,H_p'(k_2a)/J_p'(k_0a)\,H_p'(k_2a) - e^{j\pi/4}\,\frac{\sqrt{\mu_2/\epsilon_2''}}{(1-\nu^2)^{3/2}}\,J_p^2(k_0a)\,H_p(k_2a)/J_p'(k_0a)\,H_p'(k_2a) = 0$$

i.e., $$J_p(k_0^0a) - \frac{\nu_0\,\delta\nu}{\sqrt{1-\nu_0^2}}\,\frac{2\pi a}{\lambda_0}\,J_p'(k_0^0a) - e^{j\pi/4}\,\frac{\sqrt{\mu_2/\epsilon_2''}}{\sqrt{1-\nu_0^2}}\cdot J_p'(k_0^0a)\,\frac{H_p(k_2a)}{H_p'(k_2a)} - e^{j\pi/4}\sqrt{\frac{\mu_2}{\epsilon_2''}}\sqrt{1-\nu_0^2}\,J_p^2(k_0^0a)\,H_p'(k_2a)/J_p'(k_0^0a)\,H_p(k_2a) - e^{j\pi/4}\,\frac{\sqrt{\mu_2/\epsilon_2''}}{(1-\nu_0^2)^{3/2}}\,J_p^2(k_0^0a)\,H_p(k_2a)/J_p'(k_0^0a)\,H_p'(k_2a) = 0 \quad .. \quad (29)$$

where ϵ_1 has been put equal to 1 for vacuum, and k_0^0 is the value of k_0 when $\nu = \nu_0$, i.e., $k_0^0 = \dfrac{2\pi}{\lambda_0}\sqrt{1-\nu_0^2}$.

Taking $J_p(k_0^0a) = 0$, Equ. (29) gives

$$\delta\nu = -e^{j\pi/4}\sqrt{\frac{\mu_2}{\epsilon_2''}}\,\frac{\lambda_0}{2\pi a\nu_0}\,\frac{H_p(k_2a)}{H_p'(k_2a)}$$

It can be shown that in the limit as $|k_2a| \to \infty$,

$$H_p(k_2a)/H_p'(k_2a) \to j$$

Hence $$\delta\nu = \frac{\lambda_0(1-j)}{2\pi a\nu_0\sqrt{2\epsilon_2''}} \quad .. \quad .. \quad .. \quad (30)$$

where μ_2 has been put equal to 1 for a metal. Thus the effect of finite conductivity in the metal causes a small change in the phase constant which has equal real and imaginary parts. The imaginary part is to be added to that given in Equ. (27) to give the overall attenuation coefficient, while the real part represents a slight decrease in guide wavelength.

The change in phase constant for dielectric loss, given by Equ. (27), is always imaginary, so that for a propagating mode (ν_0 real) an attenuation is introduced, while for a mode which is cut off (ν_0 imaginary), $|\nu|$ is made a little greater, so that the wave decays slightly more rapidly. For the finite conductivity of the metal wall, on the other hand, $\delta\nu$ is proportional to $1 - j$ or to $j + 1$, according as ν_0 is real or imaginary. If ν_0 is imaginary, therefore, a slight real component of ν is produced, i.e., a small amount of energy does propagate in the mode, with very long wavelength. Thus for a guide whose walls are imperfect conductors, true cut-off is not realized, although it is approached sufficiently closely for practical purposes. As the frequency is reduced from infinity to zero, the value of the phase constant for a given mode moves from near unity, keeping near to, but not actually on, the real axis of ν, till it is near the origin, and then moves outwards to infinity, keeping near to, but not actually on, the imaginary axis.

Forbidden Modes

It is well known that the E_{p0} modes do not propagate in a guide of circular cross-section. Considering the HSP model, the value of ν is determined from Equ. (5), and the E_{p0} mode ($p \neq 0$) requires $k_0 = 0$, whence $\nu = 1$. Putting $k_0 = 0$, Equs (3) give $E_z \equiv 0$ and $H_z \equiv 0$. From Equs (4), all the other field components vanish. Thus only an E_{p0} mode of zero amplitude can satisfy the boundary conditions.

Let us see what further light is shed on this question by the more general approach. Equ. (29) cannot be used because this depends on the tacit assumption that $\nu_0\delta\nu \ll 1 - \nu^2$ in the expansion of $\sqrt{1-\nu^2}$. This assumption does not hold for the E_{p0} modes because ν_0, the value of ν for the HSP waveguide, is 1, so we must start from the exact characteristic Equ. (21). We take $\nu = \nu_0 + \delta\nu = 1 + \delta\nu$ in the general case of a metal with a large but finite conductivity and an air dielectric. Thus $\epsilon_2 = -j\epsilon_2''$, with $\epsilon_2'' \to \infty$, and $\epsilon_1 = 1 - j\epsilon_1''$. As in the earlier Section (The Simple Waveguide with Two Media), the contributions to $\delta\nu$ due to the non-zero value of ϵ_1'' and the finite value of ϵ_2'' can be treated separately.

Without going into details, the method, as in the earlier section, is to expand Equ. (21) and neglect all

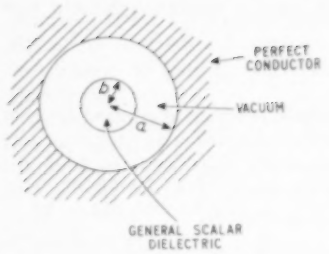

Fig. 5. *Generalized coaxial-line model*

but the lowest-order and second-order terms in $\delta\nu$, ϵ_1'', and $1/\sqrt{\epsilon_2''}$. The result is then obtained that

$$\delta\nu = \frac{-j\,\epsilon_1''}{2} + \frac{\lambda_0 p}{2\pi a}\,\frac{1-j}{\sqrt{\epsilon_2''}} \qquad .. \quad .. \quad .. \quad (31)$$

The imaginary part of $\delta\nu$, being negative, represents an attenuation, while the real part, being positive, indicates an increase in $\delta\nu$, i.e., the guide wavelength decreases and is, in fact, less than the free-space wavelength. We are accustomed to thinking of the guide wavelength as essentially greater than the free-space wavelength, but this is true only of the HSP guide.

Propagation can take place because k_0 is no longer zero, so that the field components do not all vanish identically. However, $|k_0 a|$ is small, of order $|\delta\nu|$, so that the amplitudes of E_z and H_z, given by Equs (12), will also be very small. The other field components in the air region are given by Equs (15) on putting $\mu_2 = 1$, $\epsilon_2 = 1$, and substituting for E_z and H_z. Since E_z and H_z are small, the other field components will be small. Thus only a small power flow is possible in the E_{p0} modes.

The Coaxial Line

The usual treatment of the coaxial line is to regard the metal outer and inner cylinders as perfect conductors, and the space between as a vacuum, and then to calculate the inductance and capacity per unit length. An alternative treatment is due to Clavier[2], and an account of it is given by Lamont[3]. This method adopts a quasi-static approach to the investigation of propagating waves.

It is illuminating to generalize the media of which the line is formed, and to investigate the question of wave propagation in this system In this way the point is brought out that the transmission-line mode is in fact a member of the series of waveguide modes, and that its properties of having zero z components of electric and magnetic field, and of having a phase velocity equal to that of light in free space, are not so much properties of the mode as of the system, for they only occur when the cylindrical walls are perfect conductors and the medium between them is a vacuum.

The general model of the coaxial line would take the media in the three regions $0 < r < b$, $b < r < a$, $a < r < \infty$, to be different dielectrics with complex scalar permittivities. This would be rather complicated, and for our present purposes it will be sufficient to take as model the system in which the outermost medium $(r > a)$ is a perfect conductor, the innermost $(r < b)$ is a complex scalar dielectric, and the middle medium $(b < r < a)$ is a vacuum.

We shall not derive the characteristic equation in detail; the method is analagous to that of the earlier Section (The Simple Waveguide with Two Media). We take ϵ_0 and μ_0 to be the permittivity and permeability of the vacuum, and ϵ to be the relative permittivity of the central medium $(r < b)$. ϵ is complex, and we write $\epsilon = \epsilon' - j\epsilon''$. With

$$\left.\begin{aligned} k_0 &= \omega\,\sqrt{\epsilon_0\mu_0}\,\sqrt{1-\nu^2} \\ k_1 &= \omega\,\sqrt{\epsilon_0\mu_0}\,\sqrt{\epsilon-\nu^2} \\ P &= \frac{p\nu}{b}\,\frac{k_1^2 - k_0^2}{k_0 k_1} \end{aligned}\right\} \qquad .. \quad .. \quad .. \quad (32)$$

the characteristic equation can be written

$$D_p = 0, \qquad .. \quad .. \quad .. \quad .. \quad .. \quad (33)$$

D_p being a sixth-order determinant whose elements involve k_0, k_1, P and Bessel functions of order p. All the modes of the guide can be discovered by the solution of this equation. There are two series of modes, denoted by E and H (again it must be remembered that designating a mode as E or H does not mean that H_z or E_z is identically zero). If a value is assigned to p for, say, the E series, the qth value of ν obtained is that for the E_{pq} mode. The value of q is determined by considering the value of ν as $\epsilon \to 1$, i.e., as the properties of the core material approach those of free space. Then the values of ν will approach those appropriate to the HSP guide. By comparing values of ν, then, for a given value of p, the appropriate value of q is found.

Now let us consider the core material to be a metal. This system is commonly used as a transmission line, so let us examine Equ. (33) to see if the transmission line mode is implicit in it. If the core material is a perfect conductor, $\nu = 1$. For a metal core, which approximates to a perfect conductor, we may expect ν to approximate to unity. We write $\epsilon = -j\epsilon''$, with ϵ'' very large, and $\nu^2 = 1 - \delta^2$, with δ very small. ϵ'' is real, but we put no restriction on δ. We now have

$$k_0 = \frac{2\pi}{\lambda_0}\sqrt{1-\nu^2} = \frac{2\pi\delta}{\lambda_0}$$

$$k_1 = \frac{2\pi}{\lambda_0}\sqrt{-j\epsilon''-\nu^2} = \frac{2\pi}{\lambda_0}\sqrt{\epsilon''}\,e^{-j\pi/4}$$

Thus the arguments of all the Bessel functions are either very large or very small. Approximations can then be used for the Bessel functions, and a first-order expansion made. It is found that Equ. (33) is satisfied to the first order provided $p \neq 0$.

Now, if the reduction of Equ. (33) is considered that is obtained by taking the central core to be a perfect conductor, the E_{pq} coaxial modes (p and $q \neq 0$) can be related to the E_{pq} modes of the HSP guide by letting $b \to 0$. The solution $\nu \doteq 1$ is a lower solution than that for the E_{p1} mode. The transmission line mode, therefore, corresponds to $q = 0$. It is not a single mode, but an infinite series of E_{p0} modes for all integral values of p except zero. For perfectly conducting surfaces, with vacuum between, $\nu = 1$ for all these modes, and they all have the same field patterns. For a practical coaxial line, the values of ν and the field components will differ very slightly. The transmission-line modes

will still propagate even when the materials used depart considerably from the ideal case, and we saw in the earlier Section (Forbidden Modes) that they will even propagate, although with difficulty, in the case of a metallic tube waveguide containing a dielectric, although the E_{p0} modes were not there identified with the transmission line modes.

Miscellaneous Topics

Most microwave scientists are brought up on the HSP waveguide, and learn many concepts arising out of the theory of this guide. This leads to difficulties later in life when non-HSP guides are encountered, for many of these concepts have to be discarded. We have already seen that the E_{p0} modes, which are said to be non-existent in the HSP guide, are no longer non-existent when a practical guide is considered. When a guide is reduced to a HSP guide, the field components all reduce to zero. Thus the non-existence of these modes in the HSP guide is trivial. We shall now discuss certain other topics in which some revision of conventional ideas is necessary when considering non-HSP guides.

E Modes and H Modes

In the HSP guide, for all modes, the field components have either $H_z = 0$ (E modes) or $E_z = 0$ (H modes). It is usually argued that this follows from the form of Equs (1); because each equation contains only E_z or only H_z, it is said, solutions can be obtained involving only E_z or only H_z. This does not follow at all; although the solutions obtained actually are of this type, it is not obvious from the wave equations that they will be so. Equs (9) and (10) also involve either only E_z or only H_z, and if the argument were correct, it would follow that for any mode either E_z or H_z must be zero. This would require either A and C or B and D to be zero in Equs (12) and (13). From Equ. (18), if B and D are zero, so is A, while from Equ. (19), if A and C are zero, so is B. From Equs (16) and (17), if A or B is zero, so is, respectively, C or D. Therefore E_z or H_z can only vanish if both vanish, and then from Equs (15) all the other field components vanish.

For all modes, neither E_z nor H_z vanishes. While we have shown this only for one kind of waveguide, the fact could be similarly demonstrated for any non-HSP waveguide. A more complicated example, the waveguide of circular cross-section containing a concentric rod of ferrite, has been treated by the author, and the fact demonstrated that $E_z \neq 0$ for the H_{11} mode[4]. There are still two series of modes, however, which reduce to the E modes and H modes of the HSP guide when the actual guide is reduced by changing (mathematically) the properties of its media.

It is convenient to designate the modes in a general guide as H_{pq}, E_{pq}, according to the modes in the HSP guide to which they reduce. As long as it is remembered that in general a mode designated H_{pq} does not have $E_z = 0$, there is no objection to this practice, if only it can be shown that there is a 1-1 correspondence between the modes of the non-HSP guide and those of the HSP guide. While this has not been done rigorously, it has been shown to be the case beyond all reasonable doubt for the case of a guide of circular cross-section containing a concentric ferrite rod[5], and a similar argument can be used in other cases.

Wave Impedance

In free space or an infinite homogeneous medium, the electric and magnetic fields are transverse and mutually perpendicular. The quantity E/H is defined as the wave impedance of the medium. For the HSP guide, too, a wave impedance is defined by

$$Z = E_1/H_2 = -E_2/H_1$$

where the subscripts 1, 2, refer to components of the electric and magnetic fields in the directions of the first and second co-ordinates. For the guide of circular cross-section, using cylindrical co-ordinates, this becomes

$$Z = E_r/H_\theta = -E_\theta/H_r \quad .. \quad .. \quad .. \quad .. \quad (34)$$

Using Equs (4), we thus obtain, for E modes,

$$Z = \nu \sqrt{\mu_0/\epsilon_0}, \quad .. \quad .. \quad .. \quad .. \quad .. \quad (35)$$

while for H modes

$$Z = (1/\nu) \sqrt{\mu_0/\epsilon_0} \quad .. \quad .. \quad .. \quad .. \quad (36)$$

For a given mode, then, Z is a constant, and so can be said to mean something.

Now let us consider a non-HSP guide, taking again the model of the earlier Section (The Simple Waveguide with Two Media). Using the field components given by Equs (15), and bearing in mind that neither E_z nor H_z vanishes, as was shown in the earlier Section (E Modes and H Modes), it is clear that $E_r/H_\theta \neq -E_\theta/H_r$, and that both expressions are complicated functions of position in the waveguide. In this case it is impossible to give a useful meaning to Z. The same thing has been shown to occur when ferrite is present (reference 4). It is apparent that $E_r/H_\theta = -E_\theta/H_r$, and that these quantities are constant, only if either H_z or E_z vanishes. Thus the concept of wave impedance is only valid in the HSP guide.

It might be added that a similar difficulty arises even with the HSP guide, if it is not straight[6]; although in these articles we are only considering straight waveguides.

Power Flow

The power flow in a plane electromagnetic wave is given by

$$\mathbf{P} = \mathbf{E} \times \mathbf{H}^* \quad .. \quad .. \quad .. \quad .. \quad .. \quad (37)$$

where $\mathbf{P}$ is the power flow per unit area at a given point, the flow being in the direction perpendicular to the wave front and the unit area lying in the wave front. For a wave which is not plane, this formula can be adapted by considering the power flow per unit area across an infinitesimal area. Thus for the HSP guide we write

$$d\mathbf{P} = \mathbf{E} \times \mathbf{H}^* \,.\, d\mathbf{S} \quad .. \quad .. \quad .. \quad .. \quad (38)$$

If the area $d\mathbf{S}$ is taken in the transverse plane, only the transverse components of $\mathbf{E}$ and $\mathbf{H}$ are involved, and $d\mathbf{P}$ is in the z direction. The total power in the guide is obtained by integrating Equ. (38) over the cross-section:

$$\mathbf{P} = \int\int_S \mathbf{E} \times \mathbf{H}^* \,.\, d\mathbf{S} \quad .. \quad .. \quad .. \quad .. \quad (39)$$

Further, if $d\mathbf{S}$ is taken parallel to the z axis, e.g. in a radial or cylindrical surface, $d\mathbf{P}$ is either zero or imaginary, so that no net power flows in a transverse direction.

If the guide is non-HSP, Equ. (38) no longer holds. Let us write

$$\left.\begin{aligned}\mathbf{E} &= \mathbf{E}_0\, e^{-j\beta z}\, e^{j\omega t}\\ \mathbf{H} &= \mathbf{H}_0\, e^{-j\beta z}\, e^{j\omega t}\end{aligned}\right\} \quad .. \quad .. \quad .. \quad .. \quad (40)$$

Then $\mathbf{E} \times \mathbf{H}^* = \mathbf{E}_0 \times \mathbf{H}_0$

if $\mathbf{E}_0$ and $\mathbf{H}_0$ are real. In fact, $\mathbf{E}_0 \times \mathbf{H}_0$ means $\mathbf{E} \times \mathbf{H}$ with the terms in z and t dropped, and it is to express this that $\mathbf{E} \times \mathbf{H}^*$ is written in Equs (37) and (38). But for a non-HSP guide, $\mathbf{E}_0$ and $\mathbf{H}_0$ are complex, and $\mathbf{H}^*$ no longer means $\mathbf{H}$ with $e^{-j\beta z}$, $e^{j\omega t}$, replaced by $e^{j\beta z}$, $e^{-j\omega t}$. In this case, we must write $\mathbf{E}_0 \times \mathbf{H}_0$ in Equs (38) and (39).

As an example, it is evident from Equs (15), if we bear in mind that neither E_z nor H_z is zero, that the transverse field components are complex functions of r and θ as well as of z and t. Apart from the forms of Equs (15), there is the fact that the arbitrary constants A, B, C, and D are in general complex, while the Hankel functions of Equs (13) are also complex. Another example is discussed in Reference 4—the guide of circular cross-section with ferrite. In this case, there are also real transverse components of P, indicating a flow of power round and round in loops in addition to the net flow in the z direction.

Conclusions

By treating a more general model of a simple waveguide than is usually taken, the attenuation constant due to dielectric loss and finite conductivity of the walls can be calculated, and it can be shown that E_{p0} modes exist in the waveguide of circular cross-section. The same approach can be applied in principle to the examination of 'non-existent' modes in other guides. A general approach to the coaxial line enables the transmission-line modes to be treated as waveguide modes, and it is found that they are related to the E_{p0} modes of the guide of circular cross-section.

The concept of impedence has been discussed, and shown not to be applicable to waveguides in general. The vanishing of E_z or H_z is not a general property of waveguides; it only occurs in trivial cases and is of no significance. The formula $\mathbf{E} \times \mathbf{H}^*$ for power flow requires revision for a general waveguide.

Acknowledgement

The author wishes to thank the Engineer-in-Chief of Marconi's Wireless Telegraph Company for permission to publish this article.

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MANUFACTURERS' LITERATURE

Industrial Airbrasive Unit. A 15-page booklet describing the Industrial Airbrasive Units produced by S. S. White. The airbrasive process involves directing a controlled stream of abrasive particles at a work-piece and it is designed for cutting, shaping, deburring, abrading, etc., all types of materials. With this process it is claimed that cuts as fine as 0·008 in. (or smaller, with special tips) can be made.

S. S. White Industrial Division, 10 East 40th Street, New York 16, N.Y., U.S.A.

Belling-Lee Ribbon Feeder Connector. This 2-page leaflet gives details of the Belling-Lee ribbon feeder connector and shows how a number of these can be stacked together. Ribbon feeder is a flat, flexible, insulated ribbon containing a number of parallel copper-foil-strip conductors and it is intended to replace conventional cable-forms in equipment where space is limited.

Belling & Lee Ltd., Great Cambridge Road, Enfield, Middx.

Transradio Connectors Digest. This 44-page catalogue contains detailed information and outline drawings of over 200 connectors produced by Transradio. The catalogue lists VMP, C, SM, NC, N, BNC, and 83-UHF types, most of which are produced to specific U.S. military, Joint Services, Admiralty or Air Ministry patterns.

Transradio Ltd., 138A Cromwell Road, London, S.W.7.

Acheson Dispersions. The list of Acheson dispersions has been revised and reprinted in this 12-page booklet form. Thirty-eight dispersions are classified in seven main groups and the technical data given includes carrier liquid, approximate solids content, density, diluent and flash point.

Acheson Colloids Ltd., P.O. Box 12, Prince Rock, Plymouth.

'Micro-Step' Precision Potentiometer. A 2-page leaflet describing the Cambridge 'Micro-Step' Precision Potentiometer Type 44248. It has a discrimination of one part in 2×10^7 over a single range of 2 V. The accuracy is better than $\pm 0{\cdot}001\%$ on the 2-V dial and direct readings are in steps of $0{\cdot}1\ \mu$V.

Universal Vibrograph. The 4-page leaflet gives details of the Cambridge Universal Vibrograph which consists of a clockwork recording unit for recording horizontal, vertical or rotational vibrations or deflections in structural members under load.

The instrument normally records signals from 10 c/s to 100 c/s.

Copies of these two leaflets are available from Cambridge Instrument Co. Ltd., 13 Grosvenor Place, London, S.W.1.

Shock Protection in Packaging. This 22-page booklet deals with the protection and cushioning of goods in transit. The first part discusses shock damage and its elimination and gives some basic details on the theory of cushion design. The second part includes information on the properties and types of protection materials, such as soft expanded rubber, latex-bonded animal hair and expanded polystyrene, produced by E.R.C.

Expanded Rubber Group of Companies, Mitcham Road, Croydon, Surrey.

Catalogue of Ultrasonoscopes and Accessories. This 37-page catalogue lists the products of Ultrasonoscope Co. Also it includes a section dealing with the basic principles of ultrasonic testing and descriptions of some of the techniques used for the more common applications.

Ultrasonoscope Co. (London) Ltd., Sudbourne Road, London, S.W.2.

Instruments and Controls for Research, Industry and Education. A six-page leaflet published as an abridged catalogue of products by

Servomex Controls Ltd., Crowborough, Sussex.

Telemetry, Telecontrol and Telealarm Equipment. This new publication describes some of the Pye telemetry schemes and includes paragraphs on Choice of system, Substation alarm equipment, Choice of codes for a comprehensive system, General description of telemetry equipment, Decoding and display equipment and Alarm indication for comprehensive telemetry systems.

Pye Communications Ltd., Newmarket Road, Cambridge.

Correspondence

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Electrons and Wave Mechanics

SIR,—I fear that "Quantum" must have mis-read his mile-posts when he said (p. 46 of your February issue) that an electron would have a thermal energy of 1 eV at a temperature of 10^5 °K. I take as landmark the fact that the electrons emitted from a dull-emitter cathode, at 10^3 °K, have an energy of a little less than a tenth of an electron-volt. The temperature for 1 eV is then nearer 10^4 than 10^5 °K.

If we consider the conduction electrons in a semiconductor, or in any other circumstances in which their concentration is so small that they are non-degenerate and have nearly enough a Maxwell-Boltzmann distribution of velocities, the de Broglie wavelength at room temperature can be evaluated as follows. Boltzmann's constant k has the value $1{\cdot}38 \times 10^{-16}$ erg/°K, the electron has three degrees of freedom (e.g., movement along x, y and z axes), so that if room temperature can be taken as 300 °K, the mean thermal energy of an electron, $3/2kT$, is $2{\cdot}07 \times 10^{-14}$ erg. From the relationship kinetic energy = (momentum)$^2/2 \times$ mass one can deduce the momentum as $6{\cdot}14 \times 10^{-21}$ c.g.s. units. Since Planck's constant $h = 6{\cdot}6 \times 10^{-27}$ erg-sec, the de Broglie wavelength for a free electron at room temperature is about 10^{-6} cm.

There is a further complication if the electron is in a metal where the concentration is high enough to make the 'electron gas' degenerate at room temperatures. The *mean* electron energy predicted by theory is then large, of the order of an electron-volt, although the exchangeable energy which can be detected in experiments on specific heat or electrical conduction is small. Presumably the de Broglie wavelength corresponds to the total momentum, and therefore most of the conduction electrons in a metal will have an even shorter wavelength at room temperature than the electrons which are either free or widely dispersed in a semiconductor.

One interpretation of the greater degree of tunnelling between superconductors is that the de Broglie wavelength of electrons must be greater in the superconducting state. A long wavelength implies a low value of momentum; and at fixed energy this could result from a low effective mass, since momentum = $\sqrt{\{2 \times \text{mass} \times \text{kinetic energy}\}}$. Low effective mass of an electron in the superconducting state appears quite possible since the effective mass is a function of the structure within which the electron is situated.

I would also like to suggest an alternative approach to the quantizing of six-dimensional phase-space (March 1961 issue, p. 83). Planck's quantum has the dimensions of 'action', namely $M L^2 T^{-1}$. This same group of dimensions is possessed by (a) the product energy $\times$ time (hence multiplying h by a frequency results in an energy) and (*b*) *angular* momentum. The latter leads immediately to the quantization of the orbits in the Bohr model of the atom in units of $h/2\pi$, but *linear* momentum cannot be directly quantized because it has the dimensions $M L T^{-1}$ which are not those of h. One must therefore retain the six-dimensional concept $\Delta x . \Delta y . \Delta z . \Delta p_x . \Delta p_y . \Delta p_z = h^3$. However, the number of cells which can be defined subject to a range of values of p and a specified volume $v = xyz$ will always be proportional to v. So, in general, will the number of particles to be fitted into the energy diagram. Hence the extent of degeneracy is a function of the numerical density of particles in the system, N/v, and of the temperature, but not of the volume of the system under investigation.

D. A. BELL.

Electrical Engineering Department,
The University,
Birmingham, 15.
7th March 1961.

SIR,—Dr. Bell is of course correct. I think I made the same error once before, for I see that my Kaye and Laby, page 193, now has "1 eV = 10^4 °K" pencilled large in the margin; so I haven't even got Dr. Johnson's defence.

These articles arise mainly out of my own strivings to keep abreast of developments, and I must confess that I both read and write far too quickly. I should never keep within a decade if I didn't. It is both encouraging and salutary to receive helpful criticism of this kind.

"QUANTUM".

10th March 1961.

Determinants In Theory of Harmonic Oscillators

SIR,—I can only rejoice that the neat solution proposed by Mr. S. R. Deards, p. 107, in the March issue should have arisen from recent correspondence on the above subject.

As I am a mathematician who might be said to have strayed into the field of electrical engineering, Mr. Deards' analysis of Mr. Kitai's original circuit into two simple networks in parallel-parallel never occurred to me; I only considered the equivalent circuit given later by Mr. Kitai and reproduced in my letter.

This incident indicates the importance of closer collaboration between mathematicians and electrical engineers, and of adequate specification of the real problem to be solved, which in this case was made clear by considering the original circuit direct rather than an equivalent circuit. Equivalent circuits are not unique, and tend to limit the processes of thought brought to bear upon a problem to those appropriate to the equivalent circuit so that, as here, a possible simple and neat solution is missed.

In this case, Mr. Deards' parallel-parallel analysis means that the Y-matrix is the appropriate one and its determinant is zero for oscillation. It is perhaps worth noting that if the A-matrix of an oscillatory circuit is given or easily found, the determinant of the Y-matrix of the same circuit has the value $-(a_{21}/a_{12})$ [with the sign convention of my letter, which is different from that of Mr. Deards] so that the condition of oscillation is $a_{21} = 0$.

This condition $a_{21} = 0$ is not by itself sufficient for oscillation; it could merely indicate the absence of shunt elements. A circuit with passive elements could have very small shunt admittances at certain frequencies, but these admittances could never have negative real (or resistive) parts unless the circuit contained a concealed generator capable of supplying the residual shunt loss and more, so that an oscillation was maintained and even built up. Mathematically, this is expressed by saying that $|Y| = 0$ or $a_{21} = 0$ must lead to an equation in p or d/dt having one or more roots with non-negative real parts before oscillation actually occurs.

Hitherto, we have supposed that oscillation occurred because a concealed generator over-compensated residual shunt losses; oscillation could equally well occur because a concealed generator over-compensated series losses. In this case, the condition for oscillation would be that the element a_{12} of the A-matrix was zero, or that the determinant of the Z-matrix was zero; again, zeros in p with non-negative real parts are required for actual oscillation.

J. W. HEAD.

Research Department,
British Broadcasting Corp.,
Kingswood, Surrey,
22nd February 1961.

Switching Transistors

SIR,—Mr. Barry's article (p. 442, Dec. 1960) on switching transistors contains a short section on the charge which has to be extracted or supplied to the collector–base depletion layer. I believe his presentation obscures the effects responsible. Poisson's equation —rather than capacitance, which can be introduced later—should be the starting point. The equation tells us that the width of the layer d at an abrupt junction, with uniform doping on both sides (or very heavy, though not necessarily uniform doping on one side and much more lightly uniform doping on the other) is $aV^{\frac{1}{2}}$ (with the contact p.d. absorbed into the reverse bias V), where a is a constant depending on area, dielectric constant and the doping levels. The charge on the ionized impurities of each sign in the layer is bd, where b depends on the doping levels. If we change the

voltage from V_1 to V_2 the charge changes from $Q_1 = abV_1^{\frac{1}{2}}$ to $Q_2 = \frac{1}{2}\ abV_2^{\frac{1}{2}}$. Only now do we introduce capacitance, $C = dQ/dV = \frac{1}{2}\ abV^{-\frac{1}{2}}$, enabling ab to be eliminated. If C is known to be C_0 at $V = V_0$, the charge difference $\Delta Q = Q_2 - Q_1 = 2C_0V_0^{\frac{1}{2}}\ [V_2^{\frac{1}{2}} - V_1^{\frac{1}{2}}]$ as Barry deduces. Clearly if C is measured at V_2, and $V_1^{\frac{1}{2}} \ll V_2^{\frac{1}{2}}$, $\Delta Q \simeq Q_2 = 2C_2V_2$—warning us not to fall into the error of defining a non-linear capacitance as Q/V. If the junction is linearly graded, with distance, so that Poisson's equation yields $d = a_1\ V^{\frac{1}{2}}$ (a_1 being a new constant with different dimensions from a), then $Q = \int_0^{d/2} b_1 x dx$, b being a new constant; thus $Q_1 = a_1^2\ b_1\ V^{2/3}/8$, $C = a_1^2 b_1 V^{-\frac{1}{3}}/12$ and $\Delta Q = \dfrac{3\ C_0 V_0^{\frac{1}{3}}}{2}\ [V_2^{\frac{2}{3}} - V_1^{\frac{2}{3}}]$ Making the assumption above, ΔQ is now $\simeq \dfrac{3}{2}\ C_2\ V_2$. Although all this has been implied, if not indeed said before, it may still need stressing.

J. R. TILLMAN.

P.O. Research Station,
Dollis Hill, N.W.2.
16th February 1961.

The Junction Transistor

SIR,—Mr. Achuthan's letter in your December issue contains such fallacious arguments that I feel it desirable to point them out lest their plausibility should mislead the casual reader.

He says that in his Fig. 2 for the earthed-base circuit I_0 must be less than I_{in} and, hence, that the circuit cannot oscillate. In fact, I_0 can easily be greater than I_{in}. Due to the successive phase shifts produced by the network C_1R_1, C_2R_2, C_3R_3, the current in R_3 can be shifted by 180° relative to the collector current. Providing that this current in R_3 is greater than $(1 - \alpha)I_{in}$ then I_0 is greater than I_{in}.

Similarly, in the earthed-collector circuit, Fig. 3, Mr. Achuthan is incorrect in stating that V_0 is less than V_{in}. Again, it is easy to show that the phase shifts of the CR circuits can cause V_0 to be greater than V_{in}.

As Mr. Noordanus so rightly pointed out, it is possible for a passive RC network to have a current or voltage transfer ratio greater than unity.

Finally, an equivalent circuit can only represent the behaviour of an active device with regard to its effect on an external circuit and it need not necessarily bear any relationship to the actual physical nature of the device. Hence no reasoning, such as Mr. Achuthan attempts, can possibly determine the fundamental nature of the device.

W. J. PITTS.

Applications Laboratory,
Radio & Electronic Components Division,
Associated Electrical Industries Ltd.,
Enfield, Middlesex.
25th January 1961.

SIR,—Mr. Noordanus[1] has found a fallacy in my proof[2] because he has not fully understood it. I feel that I am partly to blame for this confusion because I omitted much of the detail to reduce the size of my letter. I am well aware of the shortcomings of performing non-linear analysis in a linear manner and I avoided complications such as transient build up, hysteresis and squegging, by considering only the starting conditions of oscillation. My analysis was in fact a simple exercise in the calculus of limits. I assumed that to small signals the circuits were linear, provided $e_j < 0$ since the starting condition for oscillations is $e_j > 0$. The three circuits were then adjusted for optimum starting conditions by varying the loading G_j on each circuit as well as the coupling ratio t; the optimum value of t in each case being the value that will start oscillations under the greatest load. The three loads become identical on the verge of oscillation. In other words, the analysis was conducted on the threshold of oscillation, e_j being negative and tending, but never being equal, to zero. However, I am responsible for one misleading remark, having inferred that the conditions of optimum starting and strength of oscillations are synonymous. This was brought about by deleting what I considered to be irrelevant comments from the manuscript, because the proof in no way uses this information. These two conditions while often near together in practice (depending, of course, on hysteresis, etc.) are obviously not synonymous.

Having read the other letters on "The Junction Transistor" it seems significant that each argument is centred around oscillators. In fact the argument may be restated as follows:

> A valve or transistor oscillator can be drawn so that it appears to function in three different ways; i.e., common emitter, base, or collector. Since the three circuits are identical, the active device must operate in one mode only. What is this mode and does it correspond to any one of the three existing configurations?

Now let it be quite clear what we are discussing. This is now an exercise in logic which is mainly of academic interest and has very few practical implications. This topic has received a great response and if the existence of a basic mode is established, then we have a technique for examining any new three-terminal active device which may appear in the future. It is unfortunate that most contributors to date (both for and against a basic mode), have approached this logical poser with emphatic statements based on intuitive notions. Statements must be qualified, and the simplest way of doing this is to translate one's thoughts into equations, not words!

My contribution to this topic was, in fact, a mathematical translation and analysis of the above proposition. It was intended that the proof should be as general as possible and devoid of intuitive notions. Since evolving the proof I have attempted to invalidate it without success and I am very keen to know if it can be proved fallacious. However, it should be noted that the results of my proof predict the optimum starting condition for a Colpitts oscillator (valve or transistor) which can be substantiated by experiment. That is, neglecting frequency stability, cathode lead impedance[3,4] etc., then the two coupling capacitors $C_1 = C_2$ (and $t = \frac{1}{2}$ in my proof) should be equal.

J. R. JAMES.

Applications Laboratory,
Semiconductors Ltd.,
Swindon.
16th February 1961.

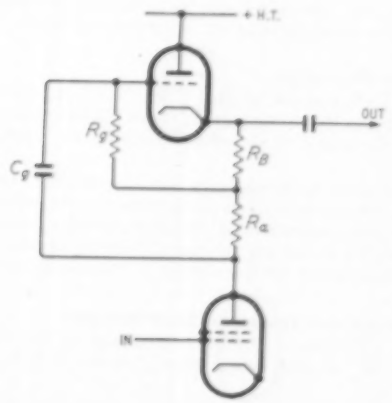
REFERENCES

[1] Ir. J. Noordanus, Letter to *Electronic Technology*, Feb. 1961, Vol. 38, No. 2, p. 69.
[2] J. R. James, Letter to *Electronic Technology*, Sept. 1960, Vol. 37, No. 9, p. 361.
[3] G. G. Gouriet, "High Stability Oscillator", *Wireless Engineer*, April 1950, Vol. 27, pp. 105–112.
[4] C. T. Kohn, "The Effect of a Cathode Impedance on the Frequency Stability of Linear Oscillators", *Proc. Inst. Radio Engrs*, Vol. 48, No. 1, pp. 80–88.

The Bootstrap Amplifier

SIR,—I was interested in the article by W. Tusting in the January issue. I beg to point out that the circuit can be much simplified.

By omitting R_d, C_d and R_b and with the temporary free end of R_a joined to the junction of R_b and R_g, we obtain the arrangement shown in the figure. I know the circuit as the 'valode'; i.e., valve-load.



A. WOMERSLEY.

Bradford,
Yorkshire.
12th January 1961.

New Books

Thermoelectric Materials and Devices

Edited by Irving B. Cadoff and Edward Miller. Pp. 344 + xiii. Reinhold Publishing Corporation, New York; Chapman and Hall, 37 Essex Street, London, W.C.2. Price 78s.

This book, the substance of lectures delivered during a course at the Department of Metallurgical Engineering at New York University, is published as one of a series on Materials Technology. This is, of course, the field where the developments of thermoelectricity are taking place. The theory of the thermoelectric circuit as a power converter was worked out about fifty years ago by Altenkirch; and it was realized that the right kind of material didn't exist. More recently, after it was found that semiconductors could be tailored to give a high figure of merit (which depends on the thermoelectric power, the resistivity, and the thermal conductivity), power generation and refrigeration became practicable propositions, and A. Ioffe's book, which was published in England in 1957, surveyed the possibilities of these materials. There is now a further quest, for materials which will operate at high-temperature differences; in the first place, so that high-temperature sources such as solar furnaces and reactors can be used for direct conversion and secondly to raise the Carnot efficiency.

The contributors to this volume survey bismuth and lead tellurides, and many other materials. Ceramics, refractory materials, ionic materials and molten salts, and liquids have been investigated; and there is also the thermionic-emission thermoelectric device. The approach of the materials technologist is to evaluate separately the parameters on which a high figure of merit depends, and several of the lectures in this book deal with measurements. On the theoretical side, there are accounts of thermal-conductivity mechanisms, and the calculation of efficiencies. Design calculations for thermoelectric generators and for Peltier cooling devices are discussed, and there is a chapter on direct conversion (by both thermoelectric and thermionic means) applied to nuclear courses, ranging from SNAP III to large-scale reactors. The basic theory is adequately covered in the earlier part of the book. G.R.N.

Transistors—Circuits and Servicing. 2nd Edn.

By B. R. A. Bettridge, M.Brit.I.R.E. Pp. 27. Published for *Wireless & Electrical Trader* by Iliffe Books Ltd., Dorset House, Stamford Street, London, S.E.1. Price 3s. (postage 4d.).

"This booklet explains in simple practical terms how transistors work, how they are used in radio circuits, and the best methods to employ when servicing equipment that uses them. The treatment is almost entirely descriptive, mathematics being avoided, and the technical level is exactly right for the service engineer."

Video Tape Recording

By Julian L. Bernstein, M.I.R.E., M.S.M.P.T.E. Pp. 268 + xi. Published in the U.S.A. by John F. Rider Publisher Inc. and in the U.K. by Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 72s.

"This definitive book presents a basic yet thorough treatment of the techniques, mechanics and circuitry used in the rapidly expanding field of video tape recording. For those having a limited technical background, the author has provided sufficient basic theory to enable them to comprehend the functioning of video tape equipment."

Avo International Transistor Data Manual

Edited by C. E. Bull. Pp. 147. Published by Avo Ltd., Avocet House, 92-96 Vauxhall Bridge Road, London, S.W.1. Price 35s. (post free).

"This Manual is intended for use in conjunction with the Avo Transistor Analyser, but the information is presented in such a manner that it forms a quick and convenient guide for general use."

Sensitometric Control in Film Making

By L. J. Wheeler, F.R.P.S. Pp. 22. A B.B.C. Engineering Division Monograph published by B.B.C. Publications, 35 Marylebone High Street, London, W.1. Price 5s.

"This Monograph describes the calibration and operation of the Type X6 Sensitometer and the Type MND Line Densitometer installed at Alexandra Palace, together with a detailed description of the system of sensitometry established to control the continuous processing of films used in the B.B.C. television news programmes.

It also discusses the subsidiary factors which must be controlled when any form of sensitometry is employed and the working tolerances which apply to each variable."

Manipulation of Thermoplastic Sheet, Rod and Tube

By J. M. J. Estevez, B.Sc., F.R.I.C., A.P.I. and D. C. Powell, B.Sc. Pp. 148. Published for the Plastics Institute by Iliffe Books Ltd., Dorset House, Stamford Street, London, S.E.1. Price 27s. 6d. (postage 10d.).

"'Manipulation of Thermoplastic Sheet, Rod and Tube' is a clear and eminently readable survey of the principal thermoplastic materials available commercially, and the methods by which they can be fabricated into useful goods. Wherever possible, a critical assessment is made of the materials and techniques available for a particular application."

Master Receiving—Picture Tube Substitution Guide Book

By H. A. Middleton. Pp. 343 + viii. Published by John F. Rider Publisher Inc., U.S.A. and distributed in the U.K. by Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 60s.

"This Guide is a revision and compilation of 'Receiving Tube Substitution Guide Book', originally published in 1950, and the four subsequent supplements. It lists 5,100 receiving tube substitutions, 825 picture tube substitutions, 325 American to European tube substitutions and 320 European to American tube substitutions."

R-L-C Components Handbook

By David Mark. Pp. 146 + v. Published by John F. Rider Publisher Inc., U.S.A., and distributed in the U.K. by Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 28s.

"It is the purpose of this book to inform the reader about the various factors of importance in selecting a resistor, capacitor, inductor or transformer and to acquaint him with the various types of available components and their special features."

Radio Astronomy

By F. Graham Smith. Pp. 264. Published by Penguin Books, Harmondsworth, Middx. Price 7s. 6d.

"Radio astronomy may be defined as 'the study of the heavenly bodies by the reception of the radio waves which they emit' . . .

This book attempts a survey of the whole of this wide subject. Inevitably it must do less than justice to parts of it and, inevitably, the parts which are treated more fully are those with which the author and the radio observatory at Cambridge have been particularly concerned."

The Training of Graduates

Pp. 34. Published by the Institution of Electrical Engineers, Savoy Place, London, W.C.2. Price 2s.

"A report of the Joint Committee on Practical Training in the Electrical Engineering Industry."

Taylor Valve Manual

Published by Taylor Electrical Instruments Ltd., Montrose Avenue, Slough, Bucks. Price 10s. (post free).

"This completely revised and up-to-date valve manual is for use with the Taylor valve tester models 45C, 45A, 46A, 47A and 45B and it includes over 5,000 valve characteristics and settings."

Mathematical Tool Kit for Engineers. 2nd Edition

By H. A. Webb, M.A., and D. G. Ashwell, M.A., Ph.D., A.M.I.C.E. Pp. 116. Longmans Green & Co. Ltd., 6 & 7 Clifford Street, London, W.1. Price 15s.

"This edition has been enlarged considerably, and contains new chapters on complex numbers, the Laplace transform (including the use of the integral inversion formula), the error function, and some little-known extensions of Macaulay's method for discontinuous particular integrals."

MEETINGS

I.E.E.

7th April. "The Conversion of Biological Data into Electrical Signals", discussion meeting to commence at 6 o'clock.

10th April. "Applications of Electrical Phenomena at Liquid Helium Temperatures", discussion to be opened by R. E. Hayes, M.Sc., Ph.D.

11th April. "Precision Instruments for Coaxial Line Measurements up to 4 Gc/s", and "The Concept of Equivalent Source E.M.F. and Equivalent Available Power in Single-Generator Calibration", by D. Woods.

12th April. "Teaching Physics to Engineers", discussion to be opened by J. Topping, M.Sc., Ph.D., D.I.C., at 6 o'clock.

17th April. "The Education of Engineers—All's Well?", discussion to be opened by the President, Sir Hamish D. MacLaren, K.B.E., C.B., D.F.C., LL.D., B.Sc.

25th April. "Simulation of Intelligence", by Professor D. M. Mackay, B.Sc., Ph.D.

26th April. "Data Transmission", by R. H. Franklin, B.Sc.(Eng.), and J. Rhodes, M.B.E., B.Sc.(Eng.).

27th April. "Medical Electronics", Fifty-second Kelvin lecture by Professor R. F. Woolmer, F.F.A.R.C.S., V.R.D.

2nd May. "Structural Properties and Applications of Evaporated Metal Films", by D. W. Pashley, Ph.D., and "Deposition of Evaporated Non-Magnetic Films", by J. R. Balmer, B.Sc.

These meetings will be held at the Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.2, and will commence at 5.30, except where otherwise stated.

Brit. I.R.E.

6th April. "Transistorized Television Receivers", discussion.

12th April. "Vibration Analysis and Testing", by D. E. Mullinger.

19th April. "Instrumentation for Obstetrics", by Dr. C. N. Smyth, M.A., B.Sc.(Eng.), M.B., B.Ch., at 6 o'clock at the University College Medical School, University Street, Gower Street, London, W.C.1.

26th April. "Electronic Counting Techniques", symposium of papers on counters employing trochatrons, transistors and magnetic film techniques.

These meetings will be held at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1, except that on 19th April, and will commence at 6 o'clock.

The Television Society

14th April. "Transparent Phosphor Screens", by Dr. D. E. N. King.

28th April. Annual General Meeting.

These meetings will commence at 7 o'clock and will be held at the Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2.

The Society of Instrument Technology

19th April. "Electronic Telephone Exchanges", by T. H. Flowers, M.B.E., B.Sc.(Eng.), to be held at 7 o'clock at Manson House, 26 Portland Place, London, W.1.

British Sound Recording Association

21st April. "Recording Vision Signals on Tape", by P. E. Axon, O.B.E., M.Sc., Ph.D., to be held at 7.15 at the Royal Society of Arts, John Adam Street, Adelphi, London, W.C.2.

The Institution of Electronics

21st April. "Modern Transistors and Their Applications", by T. D. Towers, M.B.E., M.A., B.Sc.(Econ.), at 7 o'clock in the Large Lecture Theatre, London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.

The Institute of Navigation

21st April. "Marine Radar Presentation", by S. R. Parsons and Captain F. J. Wylie, R.N., to be held at The Royal Geographical Society, 1 Kensington Gore, London, S.W.7, at 5.15.

Royal Society of Arts

26th April. "The Training of Scientists and Engineers for Industry" by Sir John Cockcroft, O.M., K.C.B., C.B.E., F.R.S., at 2.30 at the Royal Society of Arts, John Adam Street, Adelphi, London, W.C.2. Application for tickets should be made to the Secretary.

1961 COMPUTER EXHIBITION AND SYMPOSIUM

The 1961 Electronic Computer Exhibition is to be held at Olympia, London, from 3rd to 12th October.

At the same time, the Business Computer Symposium, to be held at Olympia on 4th, 5th and 6th October, will bring forward users from some of the largest and smallest firms in Britain to speak of their experience with computers.

The Exhibition and Symposium are organized jointly by the Electronic Engineering Association and by the Business Equipment Trade Association which also runs the Business Efficiency Exhibition running concurrently at Olympia.

PRIVATE INTERNATIONAL INSTRUMENTS SHOW

B. & K. Laboratories Ltd. are to hold a private exhibition from 19th to 23rd June, 1961, of the range of instruments which they distribute in the U.K.

Products from the U.K., U.S.A., Canada, Denmark, Sweden, West Germany, France, Switzerland, Holland and Italy will be shown.

The exhibition, known as the 6th International Instrument Show, is to be held at 4 Tilney Street, Park Lane, London, W.1; entrance tickets are not required.

MULLARD 'VISUAL LECTURES'

At the recent Mullard exhibition of semiconductors for industrial purposes, a new series of 10-minute filmstrip lectures were introduced. These are conventional filmstrips, but with the addition of a commentary recorded on one track of a twin-track tape. The second track carries a series of signals which can be used to control the projection of the individual frames so that picture and sound are synchronised. Six filmstrip lectures are available at present: Published Data and the Transistor; A New High Power D.C. Converter Circuit; Semiconductor Quality Control; Details of Mullard Life Testing; Semiconductor Reliability and Testing; and Transistor Environmental and Vibration Life Testing.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Deviations from nominal frequency for February 1961*

Date 1961 February	MSF 60 kc/s 1500 U.T. Parts in 10^9	Droitwich 200 kc/s 1030 U.T. Parts in 10^9
1	— 15·1	— 14
2	— 14·8	— 4
3	— 14·9	— 17
4	— 14·7	NM
5	— 14·8	NM
6	— 14·4	— 15
7	— 14·6	— 14
8	— 14·9	— 14
9	— 14·6	— 10
10	— 14·7	— 12
11	— 14·5	NM
12	— 14·5	NM
13	— 14·3	— 15
14	— 14·4	— 10
15	— 14·0	— 11
16	— 13·8	— 12
17	— 14·1	NM
18	— 14·1	NM
19	— 14·1	NM
20	— 14·3	— 11
21	— 14·1	— 12
22	— 14·5	— 9
23	— 14·7	— 8
24	— 14·6	— 7
25	— 15·0	NM
26	— 15·2	NM
27	— 14·4	— 5
28	— 14·2	— 6

* *Nominal frequency corresponds to a value of* 9 192 631 770 *c/s for the caesium* $F,m(4,0)$–$F,m(3,0)$ *transition at zero field. NM = Not measured.*

New Products

Battery-Driven Speech Recorder

A portable transistorized dictating machine measuring $8\frac{3}{4}$ in. × 6 in. × $3\frac{1}{2}$ in. and weighing $4\frac{1}{4}$ lb has been produced by Sanders Electronics. It uses standard $\frac{1}{4}$ in. magnetic tape on $3\frac{1}{2}$-in. spools and double-track recording. Dictation time on one track is 22 min, affording 44 min for both tracks.

The recorder is powered by four U2 torch batteries which provide for operation of the instrument for about 50 hr.

The microphone, which is also used as an audio reproducer, is fitted with a switch to start and stop the recorder.

As a separate unit, a foot-operated switch is provided for transcription purposes, together with stethoscopic earpieces.

W. H. Sanders (Electronics) Ltd.,
Gunnels Wood Road,
Stevenage, Herts.

Digital Volt-Ohm Meter

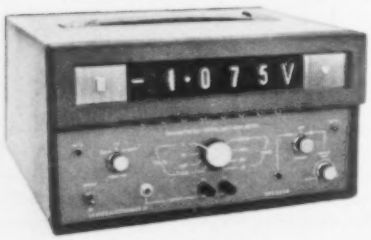
An instrument which is capable of measuring a.c. and d.c. voltages from 1 mV to 1·1 kV and resistance to 1·1 MΩ has been announced by Venner Electronics.

The measurement results are clearly displayed on in-line digital projection indicators, with automatic sign indication, and may be recorded by direct coupling to a printer and/or a reperforator control unit.

The instrument performs other digitalization functions and is suitable for operation as a single unit or for inclusion into any data logging system.

The DVOM employs no stepping switches or relays and consists of transistorized plug-in stages. It provides steady readings with fluctuating inputs and is simple to operate.

Venner Electronics Ltd.,
Kingston By-Pass,
New Maldon, Surrey.



Low-Temperature Aluminium Jointing Compound

The latest product of Eutectic Welding is a chemically-complex compound which has been designed as a high-strength low-temperature jointing material for virtually all types of aluminium and aluminium alloys. Known as Autochemic 990, it is intended for fine work such as wire and cable joining, instrument parts, small waveguides and other small fabrications.

In use, the compound is painted on where required and the joint is heated to a temperature of 400 °C; this results in a series of chain-type reactions which terminate in metallic reduction. Joints with sheer strengths of up to 9 tons/sq. in. are achieved.

Eutectic Welding Alloys, Co. Ltd.,
North Feltham Trading Estate,
Faggs Road, Feltham, Middx.

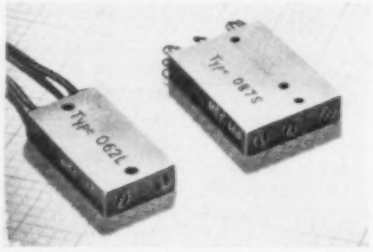
Multiple Miniature Potentiometers

Miniature pre-set type potentiometers are now being made in double and triple forms by Miniature Electronic Components.

Both the double type (Model 062) and the triple type (Model 087) are housed in a one-piece anodized aluminium case with two fixing holes, and are available with any combination of the standard ohmic values.

Each potentiometer may have a value of 10 Ω to 125 kΩ (in 17 standard values) and be fitted with solder tags or flexible leads. They will operate over the temperature range 150 °C to −55 °C, dissipating $2\frac{1}{2}$ W at 50 °C.

Miniature Electronic Components Ltd.,
The Lye, St. Johns,
Woking, Surrey.

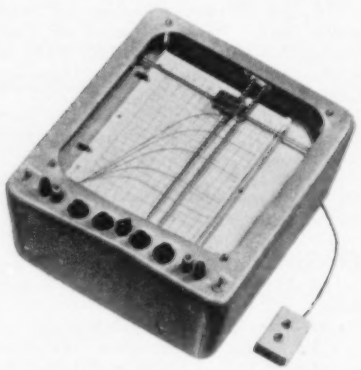


Point-Plotting X-Y Recorder

The recently-announced Houston model MR-93 X-Y recorder incorporates an electric pen-lift mechanism which allows point plotting and family curve tracing. It can also be used for conventional full-line curve tracing.

The pen lifter can be operated either manually, by using the push-button control box, or it can be designed into the test circuitry for automatic operation.

The recorder is simple in both operation and construction and uses standard $8\frac{1}{2}$ in. × 11-in. graph paper located on a flat bed. Each axis has its own amplifier, power supply, attenuator and zero control and is independent, both mechanically and electrically of the other. Amplifiers are interchangeable in 10 mV/in. and 1 mV/in. in sensitivity for either or both axes.



The accuracy is within ± 0·5% with 7·5-in./sec pen speed for both axes.

Developed by the Houston Instrument Corporation, Houston, Texas, U.S.A., this recorder is distributed in the U.K. by

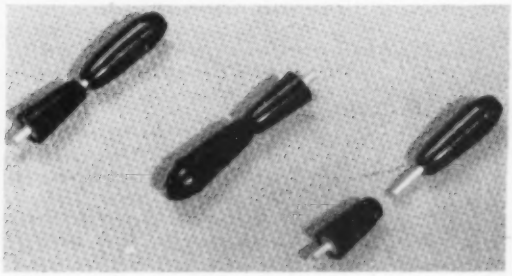
Scientific Furnishings Ltd.,
Poynton, Cheshire.

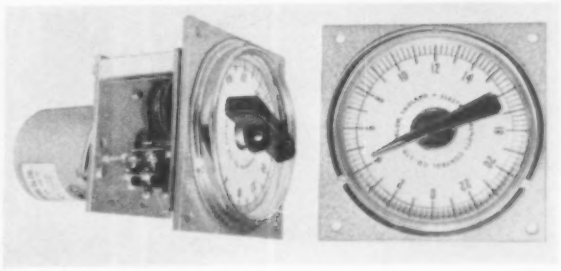
Miniature Plug and Socket

A Clix miniature wander plug and socket, particularly suited to printed-circuit mounting, has been introduced by A.E.I.

Socket-mounting pin diameter is 0·05 in., conforming to the 0·1-in. module printed-circuit technique, and sockets can be mounted with centres 0·2-in. apart. Overall diameter of the socket body is 0·19 in., pin length is $\frac{7}{64}$ in., and body length $\frac{1}{2}$ in.

The overall diameter of the plug is





0·19 in., pin diameter is 0·04 in., pin length ¼ in., and body length $\frac{9}{16}$ in. Connection to both the plug and socket is by soldering.

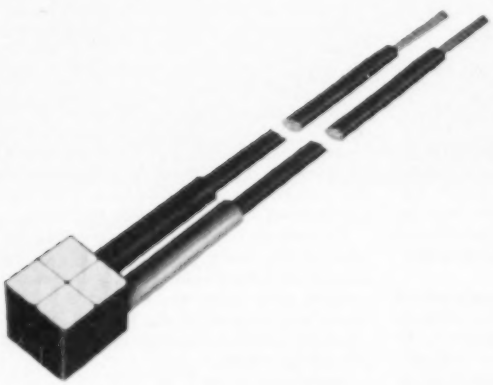
The maximum current-carrying capacity of the plug and socket is 2 A and the insulation is such that the combination will operate satisfactorily at normal television and radio h.t. voltages.

Associated Electrical Industries Ltd.,
Electronic Components Division,
155 *Charing Cross Road, London, W.C.2.*

Thermoelectric Cooling Units

The first of a range of thermoelectric cooling units has been produced by Salford Electrical.

Based on the Peltier effect, the cooling capacities of the units (i.e., the temperature-reducing capabilities) vary from 1 W for a single-junction type to more than 10 W for a large multiple junction. The electrical efficiency depends on the condition of



operation, heat sink efficiency, etc., but for a typical unit it is of the order of 30%.

The first of the range to be available is the BT.4, a four-junction unit with a cooling capacity rated nominally at 1 W. It is made in the shape of a cube measuring ½ in. × ½ in. × $\frac{5}{16}$ in. deep, and the thermal junctions are formed by a combination of p and n types of bismuth telluride semiconductor material.

Salford Electrical Instruments Ltd.,
Magnet House, Kingsway, London, W.C.2.

Hand-Set Synchronous Timer

A new hand-set timer, type NHR, which retains the front panel and fixing positions of the earlier timer, type HR, has been introduced by Elremco.

The type NHR utilizes continuously-adjustable timing cams and is fitted as standard with one or two changeover microswitches. Non-standard models with up to twelve timed circuits and twelve changeover microswitches are also available.

In operation, the timer is pre-set by hand to the required time interval and the pointer returns to zero during the timing period giving, at the same time, and indication of the progress of the timing cycle.

Switching capacity is 6 A at 230 V and time ranges are available between 0–30 sec and 0–12 days.

Electrical Remote Control Co. Ltd.,
The Fairway, Bush Fair,
Harlow, Essex.

Ultrasonic Cleaning Tank Transducer

A 1½-gallon ultrasonic cleaning tank with a 40-kc/s piezoelectric transducer mounted in the base is one of the latest additions to the Dawe range of products.

Known as Type 1160/T52, it is a welded stainless-steel assembly with internal dimensions of 8 in. × 6½ in. × 10 in. deep. The overall depth is 13¼ in. and it has a 1-in. flange at the top enabling the unit to be mounted in a bench aperture.

The unit is designed primarily for use with the Dawe type 1150, 500/125-W generator.

Dawe Instruments Ltd.,
99–101 *Uxbridge Road, London, W.5.*

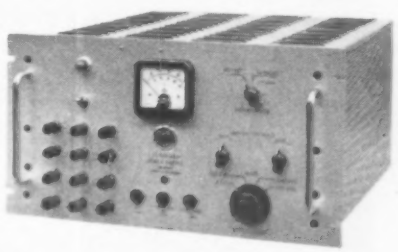
Stabilized Power-Supply Unit

A comprehensive h.t. and l.t. power-supply unit which has been designed as a self-contained equipment for bench use or for incorporation into permanent equipment has been introduced by W. G. Pye.

It is operated from the normal 110–120 V or 190–250 V, 40–60-c/s mains supply and provides positive and negative stabilized h.t. outputs and two separate floating l.t. outputs of 6·3 V a.c., 5 A; the 6·3 V output windings are centre-tapped.

The stabilized positive d.c. output is switch-controlled in four overlapping ranges (nominally 0–125 V, 125–250 V, 250–375 V and 375–500 V); within these ranges the voltage is continuously variable and on the first range a small negative voltage (−10 V) is provided. The maximum current available on all positive voltage ranges is 250 mA.

The stabilized negative d.c. output is fixed at −250 V (30 mA max.), but a



potentiometer across the negative supply provides a variable standardized output between 0 and −250 V (1 mA max.) at separate terminals.

A panel meter is provided to monitor the positive d.c. voltage and current.

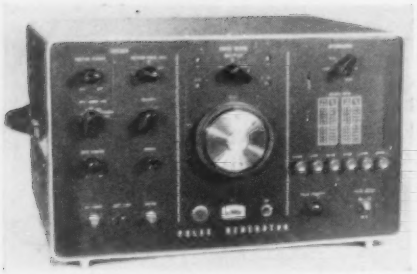
For variations of ±10% in the supply voltage the positive stabilized output does not vary by more than ±0·1% and the negative supply does not vary by more than ±0·5%. The output impedance of the positive supply is less than 0·75 Ω up to 50 kc/s and less than 1 Ω up to 1·5 kc/s for the fixed negative supply. Ripple on the positive supply is less than 3 mV r.m.s. and 1·5 mV r.m.s. for the negative supply.

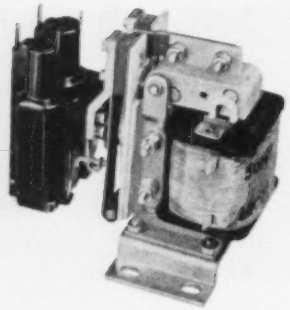
W. G. Pye & Co. Ltd.,
Granta Works, Cambridge.

Fast-Repetition Pulse Generator

The Du Mont pulse generator type 404 is an instrument providing output pulses with continuously-variable repetition rates from 10 kc/s down to a single manually-triggered pulse.

Features claimed for the instrument are rise and fall times of 0·02 and 0·025 μsec, respectively; the continuously variable pulse width from 0·05 to 100 μsec; facilities for pulse delay before or after trigger; provision of an output trigger signal to synchronize external equipment; a push-button multi-





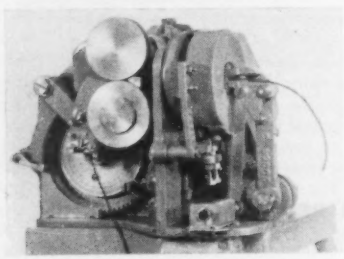
step attenuator calibrated in 0·5-dB steps to 59·5 dB, and 50-V output into a load impedance of 50 Ω.

Distributed in the U.K. by
Aveley Electric Ltd.,
Ayron Road, Aveley Industrial Estate,
South Ockendon, Essex.

Wire-Cutting, Stripping and Twisting Machine

A Swedish machine which is designed to cut multi-strand plastic-covered wire into pre-determined lengths, strip a required amount of plastic insulation off the end and twist the strands is now available in the U.K. through Vicsteels.

The machine handles wire with an outsize diameter from 0·04 in. to 0·16 in.; it cuts lengths from 2 in. to 5 in. and end-strips from 0·3 to 0·4 in. at a rate of 1500/hr. By exchanging two gear wheels it is possible to obtain cut lengths from 4 in. to 9 ft. 10 in. at a rate of 750/hr. Driven by a ½ h.p. motor, the machine measured 1 ft. 2 in. x 1 ft. 4 in. x 11 in. high and can be bench-mounted or fitted to a trolley which has a platform for



the reels of wire and a collecting trough for the cut lengths. The machine is also equipped with a comptometer.

Distributed in the U.K. by:
Vicsteels Ltd.,
Craven House,
16 Northumberland Avenue,
London, W.C.2.

Vacuum Switch

An electrical make-or-break switch in which the contacts are situated in a vacuum has been announced by B. & R. Relays for switching loads up to 2 kW (10 A max.) at voltages ranging up to 3 kV.

The basic switch, type 183, is tubular in construction with a lever actuator projecting from one end of the assembly on the axis. When supplied complete with a plastic housing, the switch is designated type 85 and a lever assembly is built into the plastic housing for operation of the switch; connection contacts are also moulded into the housing.

In addition, the plastic housing is made in such a way that up to a maximum of four of these switches can be fitted easily to the B. & R. relays types C.12 and C.62.

B. & R. Relays Ltd.,
Temple Fields, Harlow, Essex.

'Building Brick' Power Supplies

A convenient method of obtaining power supplies for transistor circuits is provided by a range of standard electronic sub-assemblies announced by Mullard Equipment.

The range consists of only four basic units and yet it is claimed that, from appropriate numbers and combinations of these, stabilized and unstabilized power supplies can be built to meet the requirements of most types of transistorized equipment. Each of the four sub-assemblies is transistorized and includes fuse and thermally-operated cut-out protection.

The four basic units comprise two unstabilized supplies with outputs of up to 1 A and 5 A respectively at 0 to 39 V, adjustable in 3-V steps; a 250 mA stabilizer with a stability factor of 200: 1 for mains variations between +10% and −15% of the nominal, and a series regulator unit (rated at ½ A, 1–30 V).

The maximum voltage from combinations incorporating the stabilizer or series regulator is 30 V. Coarse control of the output is governed by adjustable taps on the mains transformer of the unstabilized units, fine control by a pre-set potentiometer in the stabilizer.

In addition to its mains output, each of the unstabilized units has two outputs. Those of the 1-A units each provide 6, 12 or 24 V at 200 mA or 36 V at 100 mA; those of the 5-A unit each provide 6, 12 or 24 V at 1 A or 36 V at 100 mA.

Mullard Equipment Ltd.,
Torrington Place, London, W.C.1.

Accurate Angular Divider

With the angular divider which is produced by Theta Instrument Corp. U.S.A., the angular positioning of rotating devices can be measured to an accuracy of 20 seconds of arc.

In operation, the rotor of the device under test is grasped with a collet and positioned by a precision detent mechanism and the housing is secured by three clamps.

The large-angle divider is a 72-tooth hardened-steel ½ in. thick plate with a

spacing error of less than 10 seconds of arc between any two notches. Location is by a hardened-steel ball which is forced into the notches by a cantilever spring.

To span the angular interval between the 5° notches on the plate which are directly indicated on a large dial, a fine-pitch screw moves a lever arm attached to the housing adaptor and the displacement of the lever arm can be read directly from a large-scale indicator in minutes of arc.

Distributed in the U.K. by:
Livingston Laboratories Ltd.,
Retcar Street, London, N.19.

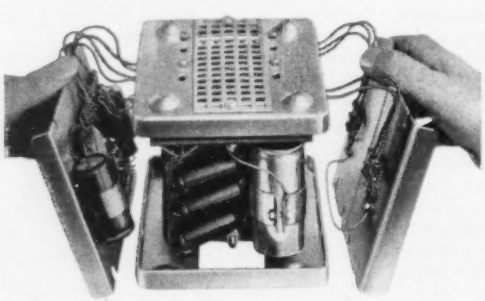
Clip-on Volt-Ammeter

The Amprobe Junior is a low-cost clip-on ammeter with two voltage-measuring ranges which has been designed to enable alternating-current measurements to be made without breaking the current circuit.

Models are available in current ranges of either 0 to 25, 0 to 50 or 0 to 100 A a.c.; all models also provide 0 to 300 and 0 to 600-V a.c. ranges. Accuracy of both the current and voltage ranges is ± 3% f.s.d. Readings are obtained on a 1·8 in. calibrated scale.

To measure current, the trigger-operated jaws of the meter are clipped on to the current-carrying conductor and, to measure voltage, plug leads are connected to the meter and clipped to the load; a button-operated switch is incorporated to change the voltage ranges.

Distributed in the U.K. by:
H. J. Baldwin & Co. Ltd.,
221 Grand Buildings,
Trafalgar Square, London, W.C.2.



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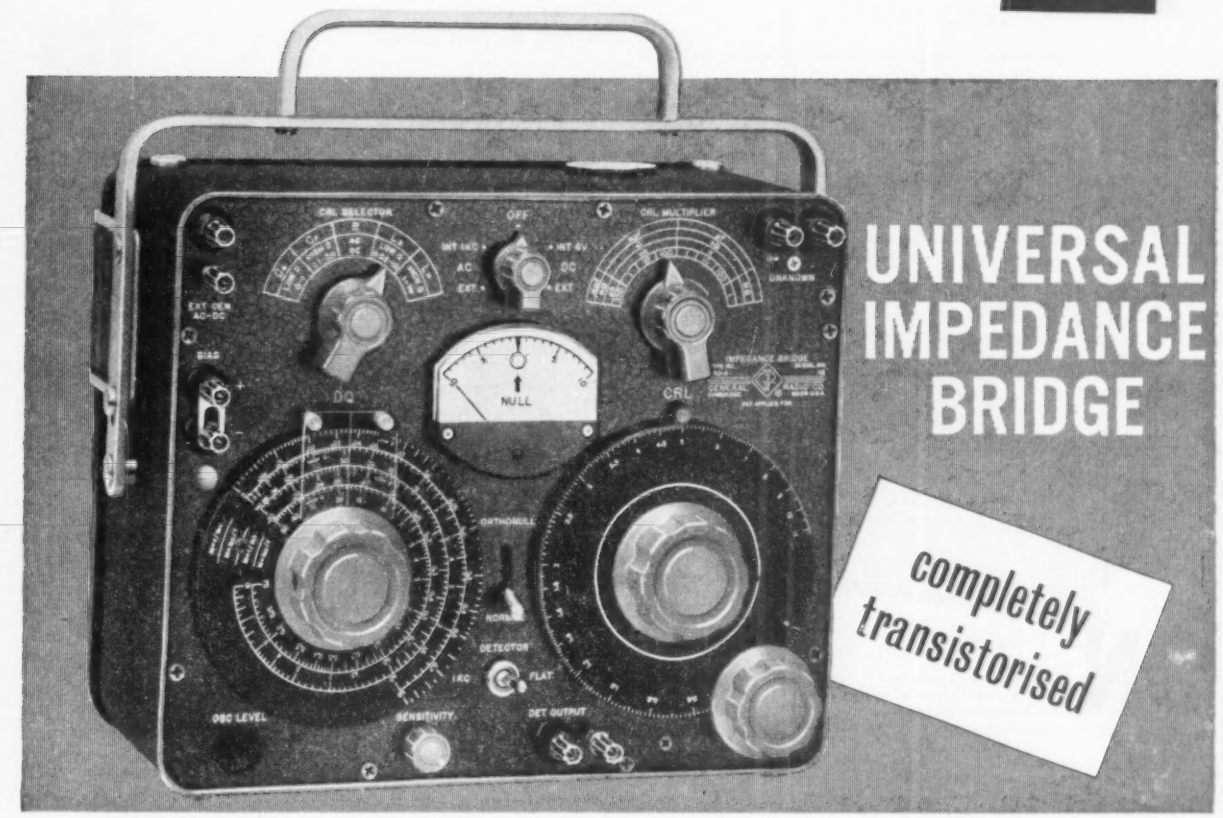

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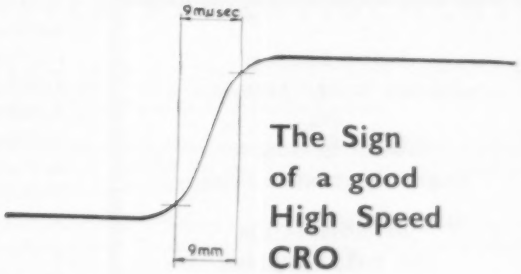

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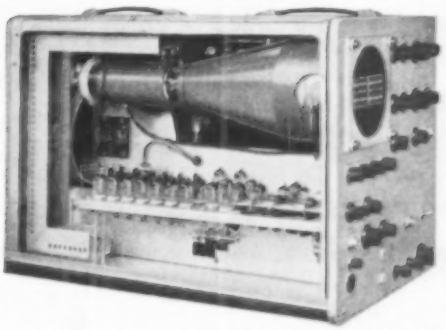

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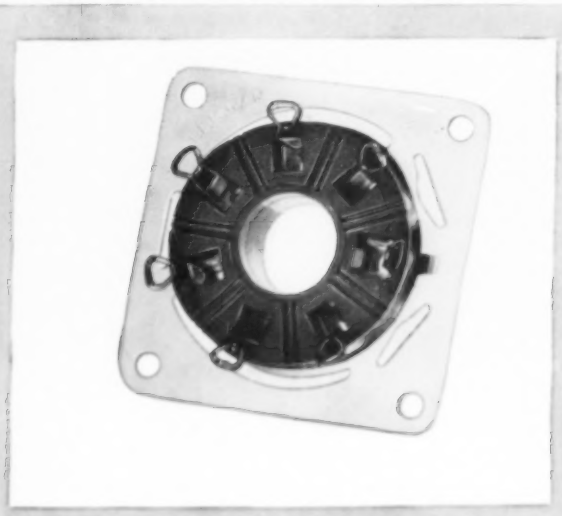

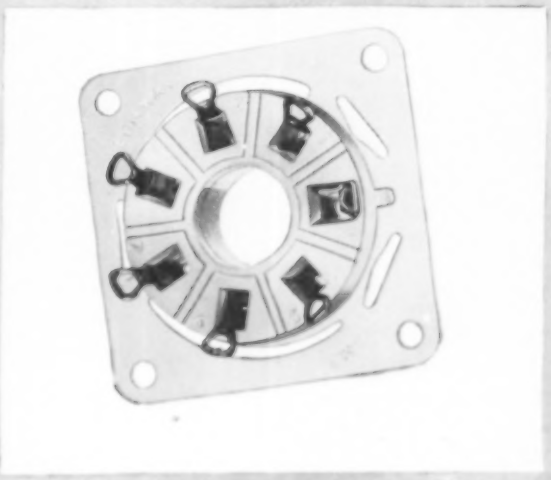

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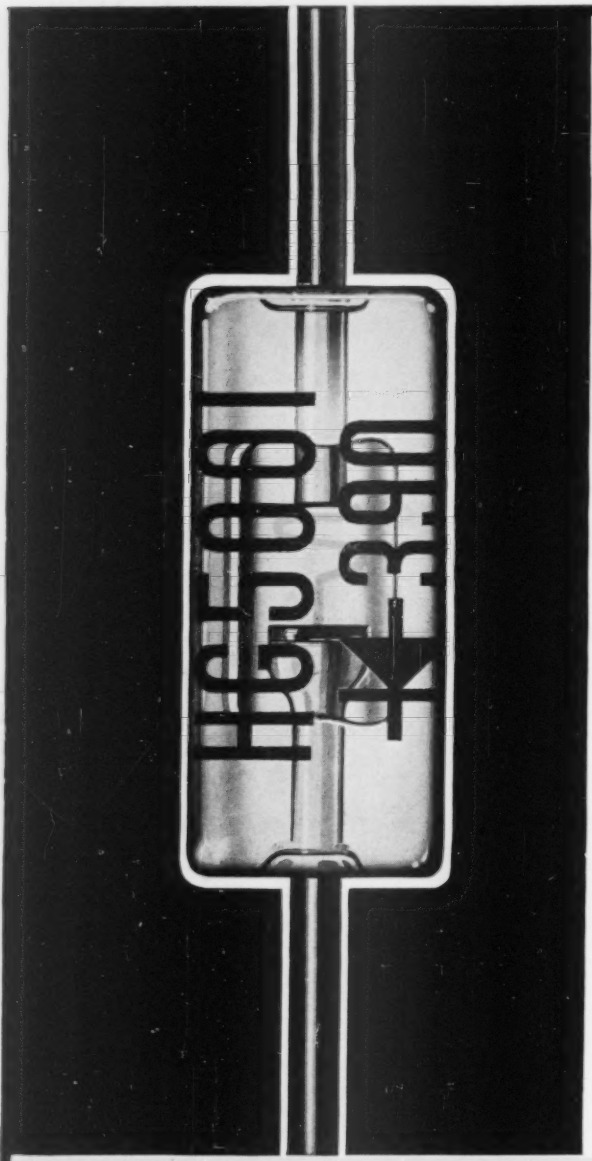

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